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A MODEL OF THE 1.6 GHz SCATTEROMETER

Job Order 75-415

Prepared By

Lockheed Electronics Company, Inc.

Systems and Services Division

Houston, Texas

Contract NAS 9-15200

For

EARTH OBSERVATIONS DIVISION

SPACE AND LIFE SCIENCES DIRECTORATE



National Aeronautics and Space Administration
LYNDON B. JOHNSON SPACE CENTER
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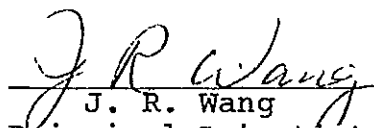
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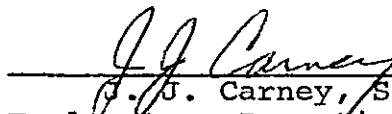
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ABSTRACT

This document is primarily concerned with the performance of the 1.6 GHz airborne scatterometer system which is used as one of several Johnson Space Center (JSC) microwave remote sensors to detect moisture content of soil. The system is analyzed with respect to its antenna pattern and coupling, the signal flow in the receiver data channels, and the errors in the signal outputs. The operational principle and the sensitivity of the system, as well as data handling are also described. No discussion is made about the interaction of the transmitted radio frequency (RF) waves with the terrain surface. The dielectric property of the terrain surface, as far as the scatterometer is concerned, is contained in the assumed forms of the functional dependence of the backscattering coefficient σ^0 on the incident angle θ .

The finite cross-polarized gains of all four 1.6 GHz scatterometer antennae are found to have profound influence on the cross-polarized backscattered signal returns. If these signals are not analyzed properly, large errors could result in the estimate of the cross-polarized σ^0 . It is also found necessary to make corrections to the variations of the aircraft parameters during data reduction in order to minimize the error in the σ^0 estimate. Finally, a few recommendations are made to improve the overall performance of the scatterometer system.

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LIST OF ACRONYMS AND ABBREVIATIONS

Acronym

AC	Alternating Current
C	Centigrade
CAD	Computation and Analysis Division
cm	Centimeter
CW	Continuous Wave
dB	deciBel
dBm	deciBels referred to 1 milliwatt in 600 ohms
DC	Direct Current
DDC	Data Distribution Center
EOD	Earth Observations Division
ESD	Engineering Systems Division
FM	Frequency Modulation
g	Gram
GDSD	Ground Data System Division
GHz	GigaHertz
GSFC	Goddard Space Flight Center
HH	Horizontal-Horizontal
HV	Horizontal-Vertical
Hz	Hertz
IR	Infrared
IRIG-A	Inter-Range Instrumentation Group Format-A
JSC	Johnson Space Center
JSME	Joint Soil Moisture Experiment

Acronym

K	Kelvin
KHz	KiloHertz
m	Meter
MHz	MegaHertz
NASA	National Aeronautics and Space Administration
PCM	Pulse Code Modulation
pf	picofarad
REC	Record
RF	Radio Frequency
RMS	Root Mean Square
SSBM	Single Side-Band Modulator
TAMU	Texas A&M University
UK	University of Kansas
VH	Vertical-Horizontal
VSWR	Voltage Standing Wave Ratio
VV	Vertical-Vertical
XMIT	Transmit

1. INTRODUCTION

Three scatterometer systems at the frequencies of 0.4, 1.6, and 13.3 GHz have recently been installed on a NASA aircraft as additional microwave sensors for the Joint Soil Moisture Experiment (JSME, ref. 1). These scatterometers at aircraft altitudes transmit RF waves downward and measure the power of the waves backscattered from a terrain. The ultimate results from the operation are expressed in terms of the backscattering coefficient σ^0 as a function of the incident angle θ of the RF waves. Since $\sigma^0(\theta)$ at a given frequency depends only on the surface properties of the terrain such as soil moisture content, surface roughness, vegetal cover, and soil type, these surface properties can be estimated (remotely sensed) by studying the magnitudes and variations of σ^0 with θ .

The scatterometer remote sensing of the soil moisture content has been performed almost exclusively on the ground surface by a truck-mounted system in the past years (ref. 2 and references therein). At satellite altitudes a short scatterometer experiment on the detection of the ground soil moisture content was performed in one of the Skylab missions (ref. 3). The results were encouraging, but not as convincing as those from the measurements of the truck-mounted system. The JSME airborne scatterometers were implemented to serve as a stepping stone from a ground-based system to a satellite platform. In addition to

being the remote sensing tools themselves, these scatterometers can also provide references and calibrations of the RF wave backscattering for the active microwave remote sensors at satellite altitudes.

The 13.3 GHz scatterometer system has been studied rather extensively in the past (refs. 4, 5, and 6). The most recent study on the system based on the new laboratory measurements of the system parameters was reported by Rosenkranz (ref. 7). However, a detail analysis has not been performed on either the 0.4 or the 1.6 GHz system. Reports on these two lower frequency systems were limited to those by the vendors and a general system description by Reid (ref. 8). It is strongly felt that at least a new, adequate analysis on these two systems is required in order to be able to assess their performances and limitations, especially after some replacements of components were made over the past years. Therefore, it is the objective of this document to provide an up-to-date analysis of the 1.6 GHz scatterometer system based on the most recent measurements of the system parameters. A similar system analysis on the 0.4 GHz scatterometer is being performed and will be reported in another document.

This document is organized into six sections. The system characteristics, the antenna and the system functional descriptions, the procedures of data handling, and the expected receiver signal level at aircraft altitudes are discussed in Section 2. The

scatterometer system analysis is given in Section 3. In this section, the data processing equations to estimate σ^0 are presented and a typical backscattered signal is traced through the receiver channels. Some coefficients appearing in the data processing equations are identified with the newly measured systems parameters in the process. The antenna pattern data are studied for the effects of side lobe and cross-polarization levels in Section 4. Section 5 gives the numerical results. These include the estimates of σ^0 and the estimates on the effects of antenna side lobe and cross-polarization levels. Finally, the major conclusions of the report and a few suggestions on the future endeavor are contained in Section 6.

Four appendices are included at the end of the document to provide a detail list of the results from the laboratory measurements on the 1.6 GHz scatterometer system parameters, the mathematical models on the mixers and the PIN diode modulators, and a listing of the computer program for the estimation of σ^0 .

2. SYSTEM DESCRIPTIONS

2.1 SYSTEM SPECIFICATIONS

The NASA/JSC 1.6 GHz airborne scatterometer is a continuous-wave Doppler radar system (refs. 8, 9 and 10), designed to measure the backscattering coefficient σ^0 as a function of the angle of incidence θ for various types of the earth's terrain. The scatterometer antenna pattern is fan-shaped which covers the θ range, along the aircraft flight path, of approximately $\pm 5^\circ$ to $\pm 60^\circ$ from the vertical. The scatterometer receives the backscattered signal at all angles of incidence simultaneously. As a result of the aircraft's forward motion, different Doppler frequency shifts are introduced in the return signals from different θ . The power of the return signal from a ground resolution cell associated with a given θ is finally retrieved by bandpass filtering at the corresponding Doppler frequency during data processing.

The overall system specifications and capabilities are given in Table 2-1. As indicated in the table, there are three functional modes of the scatterometer operation. In the first two modes of operation, the RF waves are transmitted with a single fixed polarization (either the horizontal polarization for Mode 1 or the vertical polarization for Mode 2) and the backscattered waves are received with both like and cross-polarizations. In the third mode of operation (Time-Share Mode), the transmission of the RF waves is automatically alternated between horizontal and

TABLE 2-1
THE 1.6 GHz SCATTEROMETER SYSTEM CHARACTERISTICS

TERMS

Type of System	CW Doppler
Transmit Frequency	1.6 GHz $\pm$ 10 MHz
Transmitter Power Output	1.0 Watt
Number of Antennae	Four, two transmit and two receive
Antenna Gain (two-way)	~22 dB
Antenna Beamwidth (two-way)	~10°
Antenna Along-Track Coverage	~120° fan beam
Transmitter/Receiver Isolation	~60 dB
Isolation Between Antennae	25 dB minimum
Antenna VSWR (Max.)	1.5:1
One-Way Side Lobe Level	~20 dB below main beam
Position of Peak Antenna Gain	~±50° (angles of incidence)
Maximum Skew Angle (Antenna pattern)	~±0.5°
Polarization	HH, HV, VV, VH
Functional Modes	1) H transmit V and H receive 2) V transmit V and H receive 3) Time Share, alternating between 1) and 2)
Dynamic Range of	~65 dB
Receiver Sensitivity	~-150 dBm/10 Hz
Receiver Noise Figure	~20 dB
Signal to Noise Ratio (1500 ft. altitude)	> 10 dB

vertical polarizations at a rate of 0.5 Hz. Only the first two modes of operation were applied in the Joint Soil Moisture Experiment (JSME). Between these two modes of operation, the Mode 1 turned out to be the most reliable one. The Mode 2 operation sometimes could give erroneous results possibly due to the imperfection in the original system design.

In the next few subsections, the antenna arrays, the operational principle, and the receiver sensitivity are further elaborated. The tape recording system as well as the data handling technique are also briefly described.

2.2 THE ANTENNAE

There are a total of four antennae in the 1.6 GHz scatterometer, two of them associated with the transmitter and the other two with the receiver. Each antenna consists of six dipoles in an array. The dipole orientations of the two transmitting antenna arrays are orthogonal to each other so that either the vertically or the horizontally polarized, fan-shaped beam can be generated. The two receiving antenna arrays are also oriented in the same way that both vertically and horizontally polarized RF signals can be collected. The nominal one-way, cross-track beamwidth is about 10 to 12 degrees as measured from each of the four antenna radiation patterns.

The along-track antenna beam coverage is normally from -60° in the aft to $+60^{\circ}$ in the fore directions. However, as shown in Figure 2-1, the antennae were mounted under the tail of the NASA C130 aircraft. The sloping aircraft body would certainly change

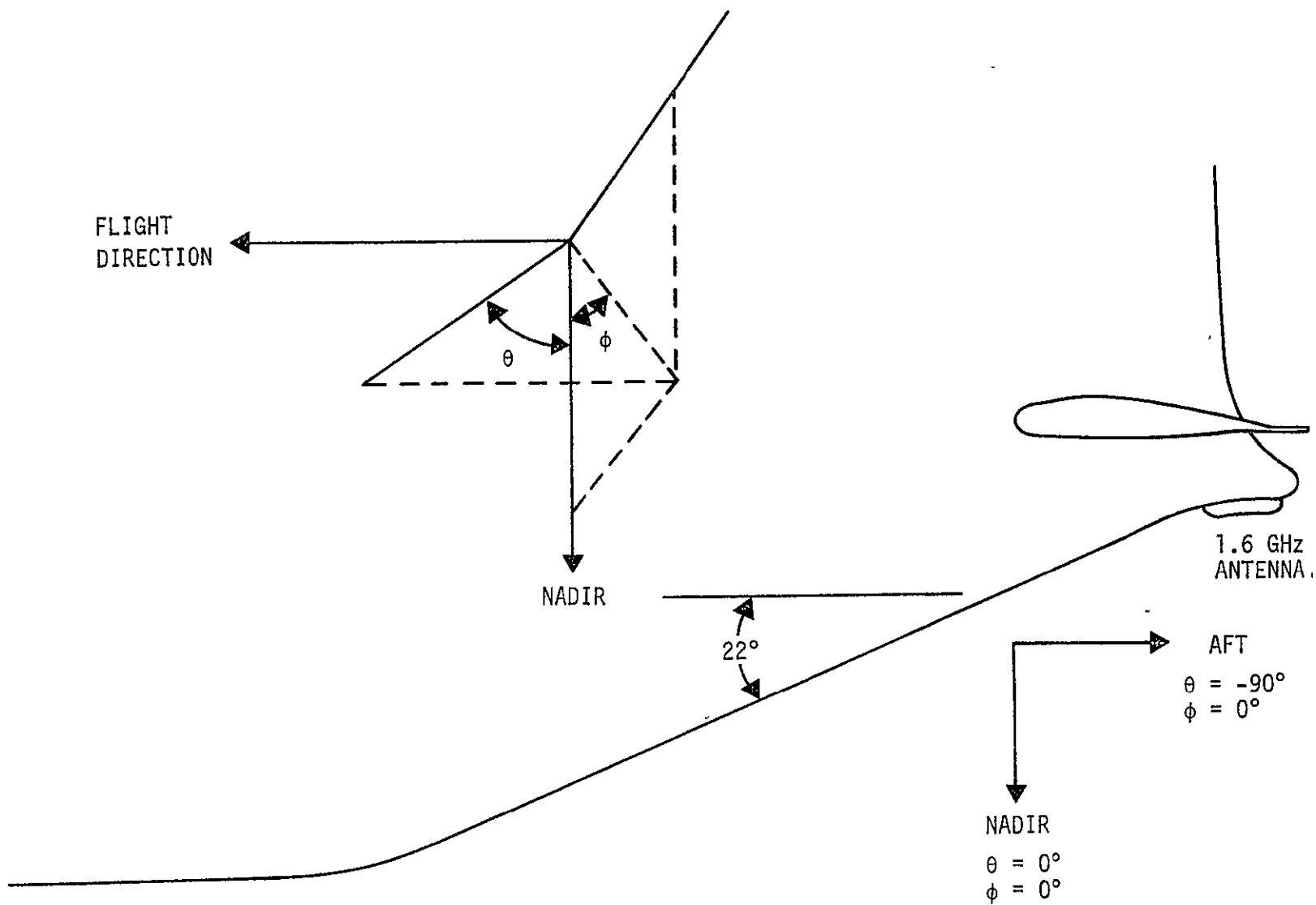


Figure 2-1.— Side view of C-130 showing coordinates for the 1.6 GHz scatterometer..

the antenna radiation patterns in the fore direction. Therefore, only the backscattered signals in the aft direction (0 to -60 degrees) can be reduced with the antenna pattern data discussed in Section 4, which were measured without a mockup to simulate the aircraft frame. The signals recorded in the fore direction may be rendered useful if the antenna gain measurements were made with a mockup simulating the aircraft frame.

Throughout this report the incident angle θ is always measured from the nadir. The plus and minus signs in θ indicate the fore and aft directions, respectively. Therefore, $\theta = -50^\circ$ implies no more than the incident angle of 50° in the aft direction.

2.3 PRINCIPLE OF OPERATION

The operational principle of the 1.6 GHz scatterometer had previously been described in several reports (reference 8, 9 and 10). For the sake of signal analysis in the later sections and the completeness of this document, it will be briefly described again in this subsection. Figure 2-2 shows the block diagram of the scatterometer system. It is essentially the same as Figure 3-1 in the report of Reid (ref. 8).

The scatterometer transmits RF power at 1.6 GHz continuously through either the horizontally or the vertically polarized antenna. For each polarization state of transmission, the backscattered signals are collected by both the horizontally and the vertically polarized receive antennae. Consequently, four distinct back-scattered signals are in order:

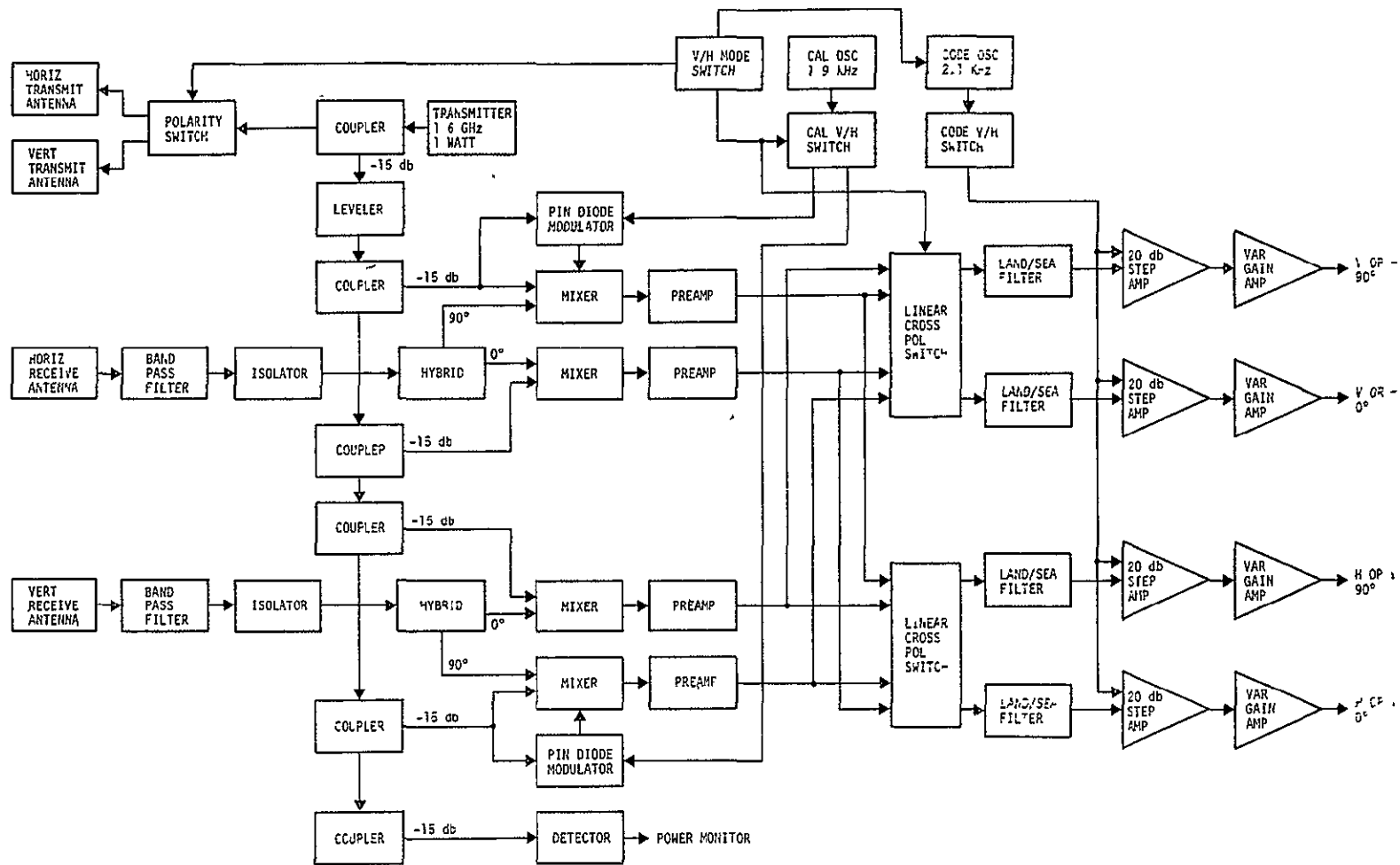


Figure 2-2.— 1.6 GHz scatterometer block diagram.

1. Horizontally polarized return when transmitting a horizontally polarized wave (HH),
2. Vertically polarized return when transmitting a horizontally polarized wave (HV),
3. Vertically polarized return when transmitting a vertically polarized wave (VV),
4. Horizontally polarized return when transmitting a vertically polarized wave (VH).

Unless otherwise specified, the description in the remaining part of this subsection refers to the horizontal receiver data channels. The operational principle for the vertical receiver data channels is essentially the same.

The backscattered signals are received at all angles of incidence (fore and aft) simultaneously. The bandpass filter, which is incorporated in each of the two receiver antenna assemblies to reduce RF interference, attenuates the received signal by a small amount ≤ 0.5 dB in the frequency range of interest. After the filter and the isolator, the signal is split into two components by a hybrid coupler. One component passes through a mixer where it is mixed with a transmitter sample. The resultant output from this mixing is a zero IF frequency spectrum data which is amplified by a series of audioamplifiers and then recorded on a tape recorder. The other component at the output of the hybrid coupler is shifted in phase by 90° and passes through another mixer where again it is mixed with a transmitter sample. In addition, a calibrate signal at 1.9 KHz from a pin diode modulator is also fed into the mixer. The resultant composite signal is translated

to audio frequency by the mixer, amplified by the audioamplifiers, and finally recorded by the same recorder. The signals from the two data channels are recorded with the same head on the adjacent channels of the magnetic tape to minimize the timing inaccuracy and skew effect. In either one of the two data channels, the fore and aft signals are folded in the mixer. As a result, the output signals from the two data channels are in phase quadrature with each other. During data reduction, these phase quadrature data are digitized and Fast Fourier transformed. Then the summation and difference of the transformed signal spectra from the two data channels are taken to recover the fore and aft Doppler signals.

The proper order of the linear and cross-polarization states of the backscattered signals at the receiver output in relation to the polarization of the transmitted waves are maintained by two linear/cross-polarization switches. The polarization switches are linked to a main control switch which determines the vertical or horizontal mode of transmitting RF waves. The main control switch is also linked to the 2.1 KHz code switch and the 1.9 KHz calibrate switch. The signal level of the coding oscillator at 2.1 KHz for a given data channel indicates the linear/cross-polarization states of the backscattered signal with respect to the transmitted RF waves. The power level of the calibrate signal depends on the state of the polarization. For example, in the JSME mission 347 carried out on October 13, 1976, the calibration constant (the K value) for the HH polarization was set at ~116 dB, while that for the HV polarization at ~131 dB.

To aid in the association of the backscattered signal at a given time to a ground location, a Zeiss camera system was installed in the aircraft to take the IR pictures of the terrain at approximately 4 sec intervals. The time of the data and picture taking enters in both the tape recording system and the picture frames. By comparing the times recorded in both the magnetic tape and the IR pictures, the backscattered signal at a given Doppler frequency (or incident angle) can be identified with a ground location, subject to a reasonable uncertainty.

2.4 THE TAPE RECORDING SYSTEM

The output of the scatterometer is recorded by a Mincom Series 110 Recorder/Reproducer System, Wideband Group I. The system is a solid state, compact, portable unit which has a bi-directional tape transport and directly interchangeable signal electronics plug-in modules for FM recording. It records data on a 1-inch magnetic tape with 14 data channels and with a carrier modulating frequency of 108 KHz. It has a bandwidth coverage from DC to 20 KHz. The signal to noise ratio is ~40 dB and the speed of the data recording is 30 inches per second. The specification of the system is given in Table 2-2.

The input level of the tape recorder is set at 1 volt RMS nominal, but is adjustable to produce full scale modulation between 0.5 and 10 volts peak to peak. The output level is 1 volt RMS nominal into a 50-ohm load. It is also adjustable from 0 to 1.5 volts RMS. The input impedance is 10,000 ohms nominal shunted by 330 picofarads maximum unbalanced to ground.

TABLE 2-2. SPECIFICATIONS OF THE TAPE RECORDER/REPRODUCER
SYSTEM USED FOR JSME (WIDEBAND GROUP I)

Magnetic Tape Size	1" in width and 14" in diameter, containing 9,200' of tape
Tape Speed	30 ips (inches per second)
Tape Speed Accuracy	0.03 ips
Time Base Error	0.6 sec at 50 KHz reference frequency
Start and Stop Time	5 sec.
% Flutter (Maximum Cumulative) at 30 ips	0.30 peak to peak at flutter bandwidth of 0.2 to 5 KHz
Center Carrier Frequency	108 KHz
Frequency Response	DC to 20 KHz
Signal to Noise Ratio	40 dB (RMS/RMS)
Input Level	1 volt RMS nominal. Adjustable for input levels between .5 & 10 volts peak-to-peak
Output Level	1 volt RMS into a 50 ohm load. Adjustable from 0 to 1.5 volts RMS
Input Impedance	10,000 ohms nominal shunted by 330 pf maximum unbalanced to ground
Output Impedance	50 ohms maximum, unbalanced to ground
Ambient Temperature Range	0° to 50°C at sea level
Thermal Drift	After 15 min. warmup, less than $\pm 0.1\%/^{\circ}\text{C}$ in full scale output voltage change, over a temperature range of 0° to 50°C
DC Linearity	Within $\pm 0.5\%$ of peak-to-peak deviation from the best zero based straight line
Total Harmonic Distortion	< 1% at 1 KHz
Humidity	Up to 95% relative humidity without condensation

The tape recorder is shared by three airborne scatterometers at the frequencies of 0.4 GHz, 1.6 GHz, and 13.3 GHz. The assignments of the channel number for various data outputs of the scatterometers are shown in Figure 2-3. The 1.6 GHz scatterometer outputs are recorded on four adjacent data channels (1, 3, 5 and 7) of the same head.

The time of data taking is recorded through IRIG Standard Time Code and Aircraft Data Acquisition System (ADAS) in channels 9 and 2, respectively. In addition to the time information, ADAS also acquires the aircraft parameters such as altitude, speed, and roll, pitch and drift angles.

The time and aircraft parameters provided by ADAS are recorded sequentially in PCM format and frequency modulated at 225 KHz. All of this information is derived during data reduction and, one way or another, is used in the data processing for the derivation of σ^0 .

2.5 RECEIVER SENSITIVITY

One way to estimate the sensitivity of the scatterometer system is by examining the signal to noise ratio at the receiver output. The signal at the receiver output can be evaluated by using the data processing equation described in section 4 and the measured system loss factor. For the convenience of a numerical estimate, Eq. (3-3) is repeated and simplified here:

$$P_r = \frac{P_t \lambda^2}{(4\pi)^3} \int_A \frac{G_t \sigma^0 G_r}{R^4} dA = \frac{P_t \lambda^2 \bar{G}_t \sigma^0 \bar{G}_r}{(4\pi)^3 R^4} A = P_t \quad (2-1)$$

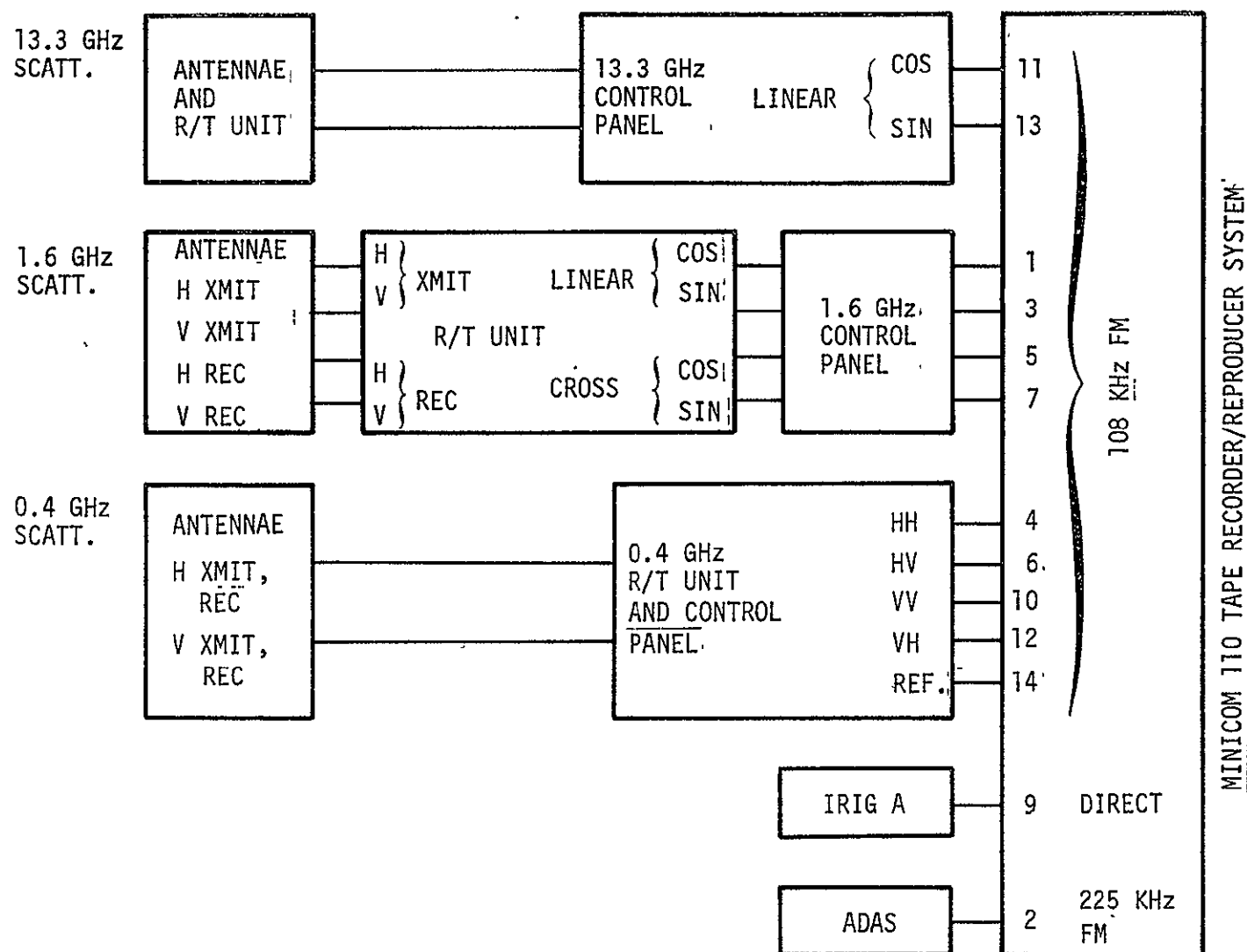


Figure 2-3.— Channel assignment of signal outputs for three scatterometers.

where a bar over a given symbol indicates the average value of that parameter over a ground resolution cell A. Besides P_t , all of the parameters on the right side of the above equation are lumped together and represented as a propagation constant C. The numerical values of these parameters were evaluated and entered in Table 2-3 for the estimate of the receiver output signal level. The aircraft altitudes and speed were assumed to be ~457m and 77 m/sec, respectively. The incident angle was taken to be 60° where low signal return was expected. The along-track ground cell length used in the area calculation was ~50m which, at 60° incident angle and 77 m/sec aircraft speed, corresponded to a bandwidth of ~8 Hz. The cross-track beamwidth as obtained from Figure 4-6 was $\sim 16.6^\circ$. The antenna feed line losses listed in the table were the measured values for the case of HH polarization (Appendix A). The σ° value at 60° incident angle was derived from Figure 3-6; the smooth water value was used for low signal return. All of the remaining parameters appeared in Eq. (2-1) are self-explanatory.

The sum (in dB) of the radiated power and the propagation constant gives the received power at the antenna terminal. After taking into account the loss due to the receive antenna feed line, the receiver signal output is -117.9 dBm. The amplification factor and the rolloff function were not entered in this calculation because both of them applied to the signal and noise equally and cancelled out when signal to noise ratio was taken.

TABLE 2-3. THE 1.6 GHz SCATTEROMETER SIGNAL TO NOISE RATIO

PARAMETERS	POWER LEVEL
Output of the Transmitter	+ 30 dBm
Transmit Antenna Feed Line Loss	- .05 dB
RADIATED POWER	+ 29.5 dBm
λ^2 (Wavelength 0.1875 m)	- 14.6 dB
$G_t G_r$ (Gain at $\theta = 60^\circ$)	+ 24.6 dB
σ° (at $\theta = 60^\circ$)	- 30.0 dB
$(1/4\pi)^3$	- 33.0 dB
$(1/R)^4$ (Range)	- 119.3 dB
A (Area Element)	+ 26.8 dB
PROPAGATION CONSTANT C	- 145.5 dB
RECEIVED POWER AT ANTENNA TERMINAL	- 116.0 dBm
RECEIVE ANTENNA FEED LINE LOSS	- 1.9 dB
RECEIVER SIGNAL OUTPUT	- 117.9 dBm
kT	- 174.0 dBm
BANDWIDTH CORRESPONDING TO RESOLUTION CELL A	+ 9.2 dB
RECEIVER NOISE FIGURE	20.0 dB
RECEIVER NOISE POWER	- 144.8 dBm
SIGNAL TO NOISE RATIO	+ 26.9 dB

For the calculation of the receiver noise level, the ambient temperature was assumed to be $\sim 290^{\circ}\text{K}$. The Boltzman's constant k is 1.38×10^{-23} joule/ $^{\circ}\text{K}$. The noise figure of 20 dB was obtained from the report by Ryan Aeronautical Company (ref. 10). Summing all of these factors, the receiver noise output over the 8 Hz bandwidth is estimated to be -144.8 dBm, which results in a signal to noise ratio of ~ 27 dB.

For operation over land or for small angle of incidence, the signal to noise ratio is expected to improve. For example, σ^0 over land could be ~ -3 dB (see Figure 3-5.) instead of -30 dB for water. The signal to noise ratio in that case would be ~ 55 dB rather than ~ 27 dB. However, the signal to noise ratio of the tape recorder system described in the previous subsection is ~ 40 dB, which represents the best value attainable with the 1.6 GHz scatterometer and data recording system.

2.6 THE SCATTEROMETER DATA HANDLING

A fair amount of data handling is required before the processed scatterometer data can be delivered to the principal investigators for studies and interpretations. Figure 2-4 gives a flow chart of the scatterometer data handling. The analog tape, which contains the data for all three scatterometer systems (at frequencies of 0.4 GHz, 1.6 GHz, and 13.3 GHz) from a given mission, is duplicated at Ground Data System Division (GDSD). The aircraft parameters such as roll, drift, pitch, altitude, and speed are also extracted and prepared in a tabular form at GDSD. The copies

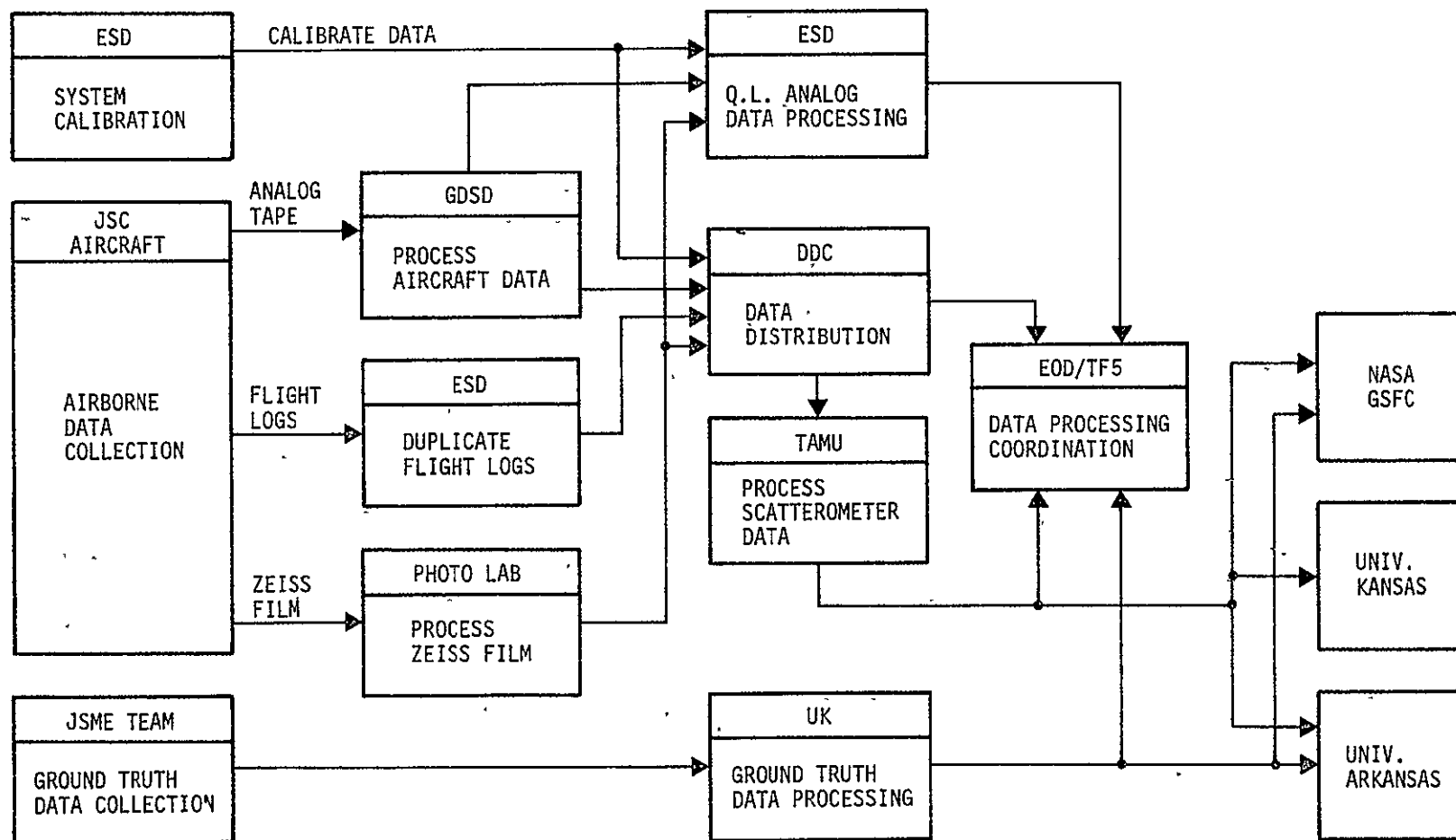


Figure 2-4.— Scatterometer data handling flow chart.

of the analog tape and the tables of aircraft parameters are delivered to the Engineering Systems Division (ESD) and the Data Distribution Center (DDC), respectively for a quick-look processing and for distribution. The results from the quick-look analog processing are examined for data quality of the mission at the SF5 branch of the Earth Observations Division (EOD/SF5).

In addition to the recording of the scatterometer data and the aircraft parameters, the flight logs and the Zeiss IR film are also generated in the mission. The flight logs contain information on the occurrence of events for the entire mission, while the IR film shows the ground scene of all the aircraft flight lines of interest. The IR film is processed and copies made at the JSC Photo Laboratory, and the flight logs are duplicated at ESD. Copies of both the IR film and the flight logs are delivered to DDC for further distribution. ESD also receives a copy of the IR film to aid in the quick-look data processing.

Generally, a system calibration is performed on the scatterometers in ESD before a major mission takes place. The calibrate constants are needed in both quick-look and final data processings at ESD and Texas A&M University (TAMU), respectively. TAMU receives a copy each of the analog tape, the aircraft parameters, the flight logs, and the calibrate constants to do the data processing with a digital method. The final outputs from the TAMU processor are plots and/or listings of the time history of the backscattering coefficient σ^0 . These outputs are distributed to

the principal investigators at NASA/GSFC, University of Kansas, and the University of Arkansas for further studies. EOD/SF5 also receives a copy for bookkeeping and evaluation purposes.

Finally, the ground truth data collection is performed jointly by all the parties involved in the JSME mission. The collected data are processed and organized at the University of Kansas. The final ground truth data outputs are tabulated and distributed to all the principal investigators involved and to EOD/SF5 at JSC.

3. SYSTEM ANALYSIS

3.1 DATA PROCESSING EQUATIONS

The power P_r reflected from a target and collected by a receiving antenna can be expressed in terms of the radar parameters as (ref. 11):

$$P_r = \frac{P_t G_t G_r \sigma \lambda^2}{(4\pi)^3 R^4} \quad (3-1)$$

where

P_t = power transmitted by the radar in watts

G_t = gain of the transmitting antenna in the direction of the target

G_r = gain of the receiving antenna in the direction of the target

σ = radar cross-section of the target

λ = radar wavelength

R = distance between the radar and the target

For remote sensing of a surface terrain by an airborne radar system such as the NASA 1.6 GHz scatterometer, it is more appropriate to use the backscattering coefficient σ^0 (differential cross-section per unit area) rather than σ . σ^0 and σ are related through the differential area element dA by

$$d\sigma = \sigma^0 dA \quad (3-2)$$

In terms of this definition Eq. (3-1) becomes

$$P_r = \int_A \frac{P_t G_t G_r \sigma^0 \lambda^2}{(4\pi)^3 R^4} dA \quad (3-3)$$

The main objective of the scatterometer is to measure P_r from a given terrain and over a wide range of incident angle θ . Then σ^0 as a function of θ can be derived and the properties of the terrain studied.

The power $P_r(\theta)$ observed at a given central Doppler frequency $f_D(\theta)$ and within a bandwidth $\Delta f_a(\theta)$ is uniquely related to a ground cell A, as shown in Figure 3-1, through the relation

$$f_d(\theta) = \frac{2V}{\lambda} \sin \theta \quad (3-4)$$

and

$$\Delta f_d(\theta) = \frac{2V}{\lambda} \cos \theta \Delta \theta \quad (3-5)$$

where λ is the wavelength of the transmitted RF waves and V is the speed of the aircraft. The two isodopplers indicated in Figure 3-1 are defined by the two Doppler frequencies at $f_d + \Delta f_d$ and $f_d - \Delta f_d$. For numerical computations described in section 5, the ground cell $A(\theta)$ is further divided into N elements and Eq. (3-3) can be rewritten as

$$P_r(\theta) = \frac{P_t \lambda^2}{(4\pi)^3} \sum_{i=1}^N \frac{G_{t1}(\theta) G_{ri}(\theta) \sigma_i^0(\theta)}{R_1^4(\theta)} \Delta A_i \quad (3-6)$$

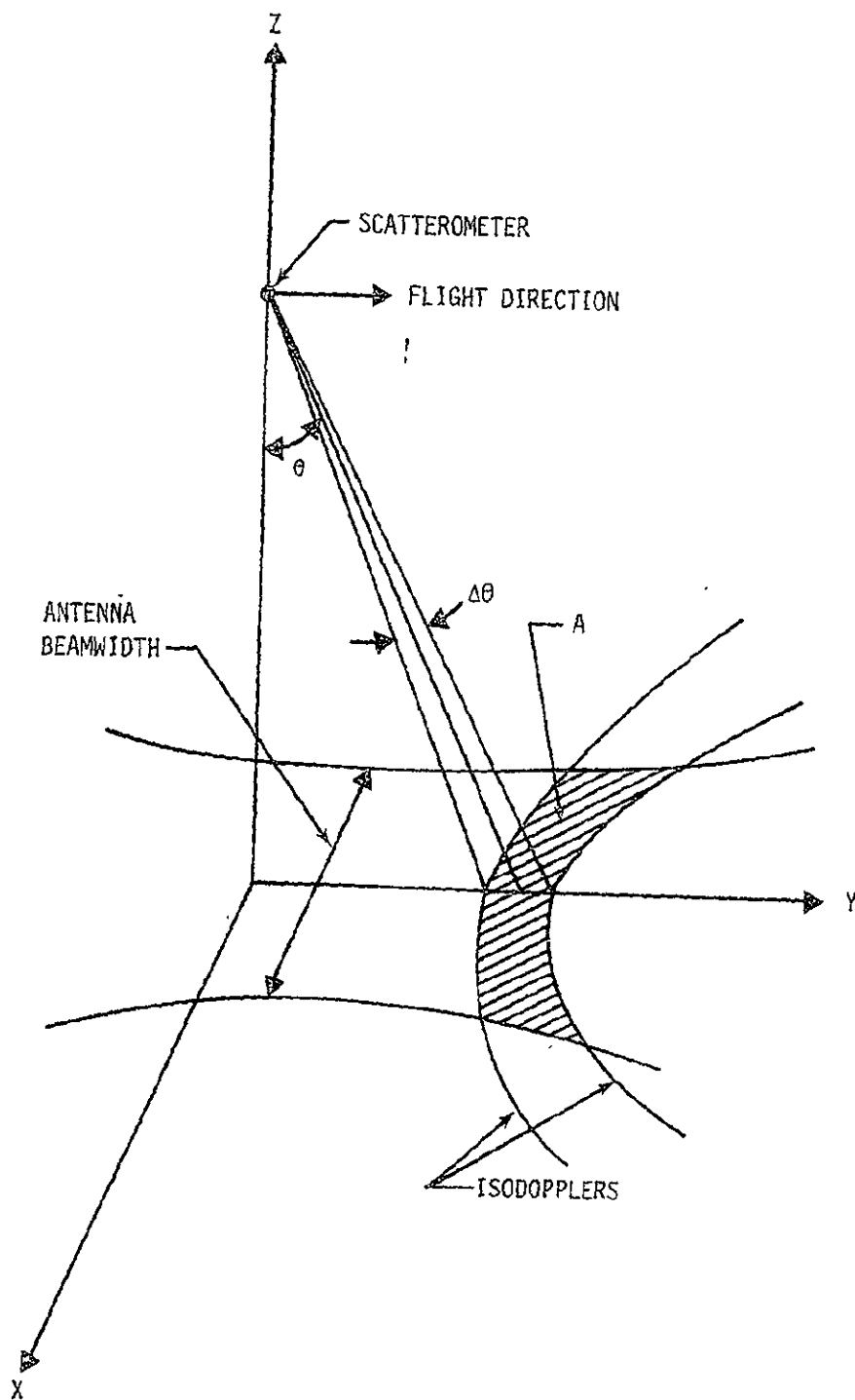


Figure 3-1.-- The resolution cell as defined by Doppler frequency bandwidth and antenna beamwidth.

The power density $W_r(\theta)$ is simply

$$W_r(\theta) = \frac{P_r(\theta)}{\Delta f_d(\theta)} = \frac{P_t \lambda^2}{\Delta f_d(\theta) (4\pi)^3} \sum_{i=1}^N \frac{G_{ti}(\theta) G_{ri}(\theta) \sigma_i^0(\theta)}{R_i^4(\theta)} \Delta A_i \quad (3-7)$$

Several parameters have been omitted in the above discussion. They should be considered before the final receiver output is obtained and interpreted. First, the attenuation of the RF power due to the two-way path of propagation is not included. The magnitude of this attenuation, in absence of heavy precipitation, depends on the amounts of oxygen and water vapor along the propagation path. Assuming that the aircraft is cruising at an altitude of 40,000 feet, the total two-way power attenuation at 60° incident angle is found to be only ~ 0.14 dB at 1.6 GHz (ref. 12). For most of the mission carried out by the C130 aircraft so far, the cruising altitude was $\sim 1,500$ feet. Therefore, the loss due to atmospheric attenuation can be completely neglected. Secondly, Equations (1), (3), and (6) apply to each of the four polarization combinations described in section 2. Thus, P_r , P_t , G_r , and G_t can be in either horizontal or vertical polarization state. Depending on the polarization state of P_r (or G_r) and P_t (or G_t), four different backscattering coefficients can be derived from the operation of the scatterometer, namely, σ_{HH}^0 , σ_{VV}^0 , σ_{VH}^0 , and σ_{HV}^0 . The subscripts to indicate the polarization state of P_r , P_t , G_r , G_t , and σ^0 are not explicitly written out in the expression above. Thirdly, P_r as given in Eq. (3-6) is the power of the back-scattered signal at the receiver input. As the signal passes

through the receiver data channels, it experiences losses as well as gains due to the presence of many subsystems. These factors will have to be taken into account in order to correctly interpret the observed data.

To derive the backscattering coefficient σ^0 as a function of incident angle θ from Eq. (3-7) requires an iteration process. First, a functional dependence of σ^0 on θ has to be assumed. Then the power density $W_r(\theta)$ of the backscattered signal is calculated from Eq. (3-7) for a series of θ and $\Delta\theta$. The calculated $W_r(\theta)$ is compared with the power spectrum derived from observations. If the power spectra from computation and from observation do not match well over the incident angles of interest, another form of $\sigma^0(\theta)$ is assumed and the procedure repeated. The final functional dependence of σ^0 on θ is ascertained when best match between the computational and the observational power spectra is obtained.

A simple, direct, and most used approach of determining σ^0 as a function of θ is to first associate a ground resolution cell $A(\theta)$ with a single average value of $\sigma^0(\theta)$, as shown in Figure 3-2, and then express $\sigma^0(\theta)$ in terms of all measurable parameters. From this figure,

$$\begin{aligned}
 d_1 &= h\Delta\phi \\
 \text{and} \quad d_2 &= \frac{R\Delta\phi}{\cos \theta} \\
 A(\theta) &= d_1 d_2 = \frac{hR\Delta\phi\Delta\theta}{\cos \theta}
 \end{aligned} \tag{3-8}$$

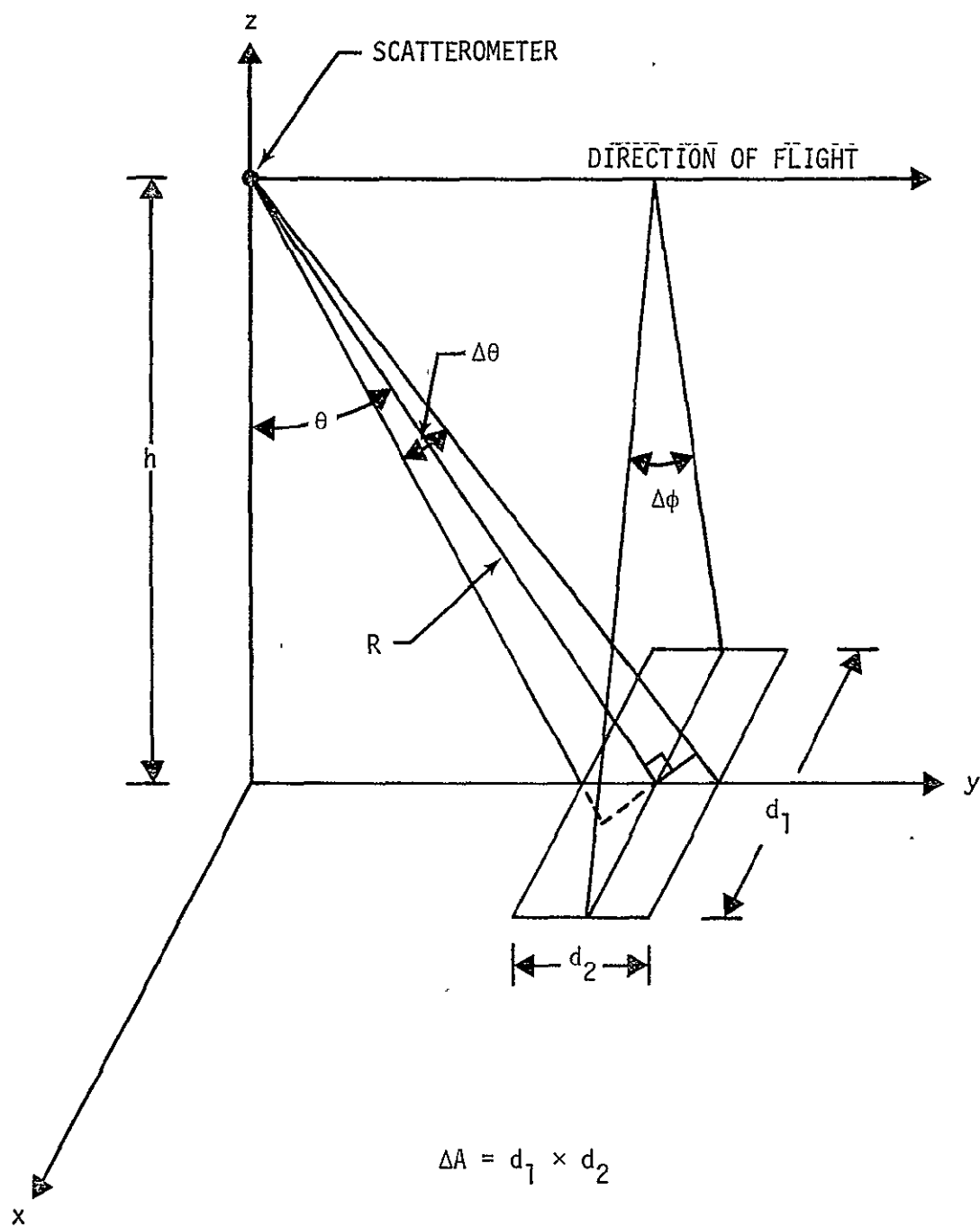


Figure 3-2.— Geometry of a scatterometer ground resolution cell.

Substituting the expression for $\Delta\theta$ in Eq. (3-5) into Eq. (3-8), the area of the ground cell becomes

$$A(\theta) = \frac{hR\lambda\Delta\phi\Delta fd}{2V \cos^2\theta} \quad (3-9)$$

and the backscattered power from this area is

$$P_r(\theta) = \frac{P_t G_t(\theta) G_r(\theta) \sigma^0(\theta) \lambda^3 \Delta\phi\Delta fd}{(4\pi)^3 2VhR} \quad (3-10)$$

Expressed in terms of $\sigma^0(\theta)$, Eq. (3-10) becomes

$$\sigma^0(\theta) = \frac{(4\pi)^3 2VhR W_r(\theta)}{P_r G_t(\theta) G_r(\theta) \lambda^3 \Delta\phi} \quad (3-11)$$

where the aircraft altitude $h = R \cos \theta$ and Eq. (3-7) was used.

Both approaches for the determination of σ^0 versus θ relationship are attempted in section 5 where the numerical calculation is performed. A comparison is made on the results from both approaches.

The backscattered signal in the receiver data channels is further analyzed and the backscattering coefficient σ^0 are discussed in the following subsections. The antenna gain pattern, side lobe level and coupling effect are discussed in detail in section 4.

3.2 RECEIVER SIGNAL ANALYSIS

To develop a comprehensive mathematical model for the signal

response of the scatterometer system requires a complete knowledge on the characteristics of each of the many subsystems , shown in the block diagram of Figure 2-3. These characteristics include the gains, bandwidths, and noise figures of all amplifiers, the gains and bandwidths of all filters, to name just a few. To make new measurements on all of these parameters may be costly and time consuming, while to develop a mathematical model based on old values of these parameters may not be appropriate, especially when some of the components were replaced over the past years. Consequently, a different approach is attempted here to derive a practical transfer function of the whole system, with the aid of some new measurements on a few key system parameters.

The approach follows closely with those attempted for the 13.3 GHz airborne scatterometer system by Krishen et al., (ref. 4) and by Bradley (ref 5). The backscattered signal is followed through all major subsystems. The final outputs from the receivers contain factors such as the gains of the amplifiers, the coefficients of the PIN diode modulators, and the gains of mixers and filters, which are combined and identified with the key system parameters recently measured in the laboratory (see Appendix A). Without loss of generality, only the horizontally-transmit and horizontally-receive mode of operation is described in the following paragraphs. Other modes of operation can be described in the same way.

Assuming that the continuous wave at frequency ω_0 at the horizontal transmitter port is represented by

$$V_t = V_o \cos \omega_0 t \quad (3-12)$$

The voltage at the input of the horizontal receiver is

$$V_r = C_f \cos [(\omega_0 + \omega_d)t + x_f] + C_a \cos [(\omega_0 - \omega_d)t + x_a] \quad (3-13)$$

where

C_f = amplitude of the received signal in fore direction, in volts

C_a = amplitude of the received signal in aft direction, in volts

ω_d = the Doppler shift corresponding to fore and aft beam at a particular angle of incidence, in radians/sec

x_f = the phase shift introduced by the path length and the rough surface in the fore beam, in radians

x_a = the phase shift introduced by the path length and the rough surface in the aft beam, in radians

The direct leakage of the signal from transmitter to receiver at frequency ω_0 is not included in Eq. (3-13) for reason given in section 4.

The hybrid coupler in the receiver channel divides the input signal equally into two parts, one without phase shift and the other with a 90° phase shift. The output signals from the

hybrid coupler can be written as

$$V_{1c} = \frac{K_1}{\sqrt{2}} \left[C_f \cos ((\omega_o + \omega_d)t + x_f) + C_a \cos ((\omega_o - \omega_d)t + x_a) \right] \quad (3-14a)$$

$$V_{2c} = \frac{K_2}{\sqrt{2}} \left[-C_f \sin ((\omega_o + \omega_d)t + x_f) - C_a \sin ((\omega_o - \omega_d)t + x_a) \right] \quad (3-14b)$$

where V_{1c} and V_{2c} are the output signal voltages for the channel (channel 1) without phase shift and the channel (channel 2) with 90° phase shift, respectively. k_1 and k_2 take care of the loss in the isolator and the possible unbalance in the signal power division of the hybrid coupler. There exists an asymmetry between channels 1 and 2 as shown in Figure 3-3. In channel 1, the back-scattered signal from the hybrid output is directly mixed with the transmitter sample signal in a mixer. In channel 2, however, the transmitter sample signal is coupled to and modulated by a 1.9 KHz calibrate signal through a PIN diode modulator. The combined signal from the PIN diode modulator output is coupled with the backscattered signal and the resultant signal is mixed with the transmitter sample in the channel 2 mixer. The voltage of the transmitter sample signal is

$$V_s = \rho V_o \cos \omega_o t \quad (3-15)$$

and the transfer characteristic of the PIN diode modulator can be written as (see Appendix B):

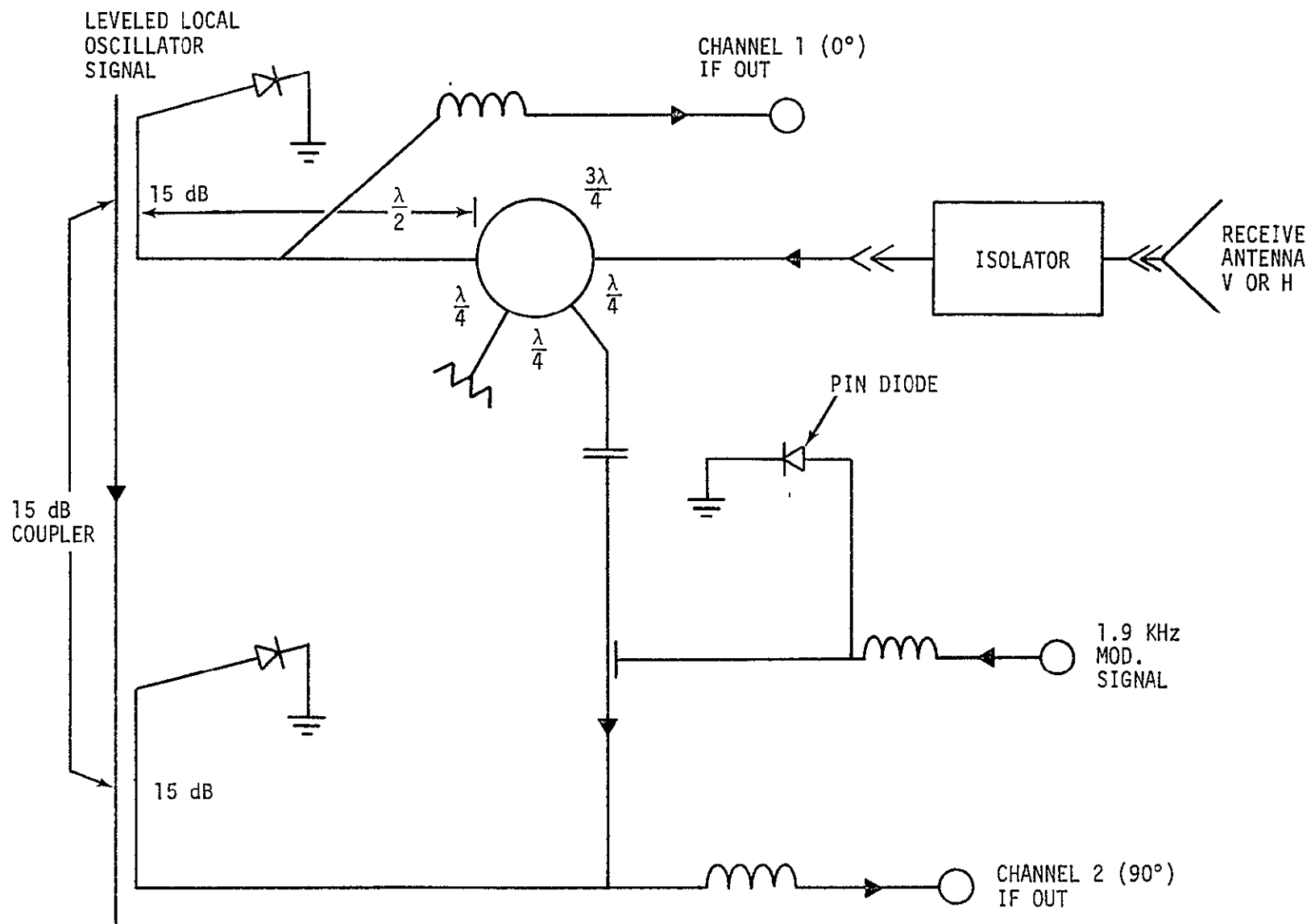


Figure 3-3.— 1.6 GHz Scatterometer Microwave Stripline Chart.

$$I_b = \alpha(1 + \gamma \cos \omega_c t) \quad (3-16)$$

Here ρ gives the magnitude of the amplitude attenuation in the transmitter signal. ω_c is the angular frequency of the calibrate signal. α and $\alpha\gamma$ ($\gamma \ll 1$) are the amplitudes of the DC and AC bias currents, respectively.

After the mixers, the output signal voltages from both channels are (see Appendix C):

$$V_{1m} = \frac{k_1 \beta_1 \rho_1 V_o}{\sqrt{2}} \left[C_f \cos(\omega_d t + x_f) + C_a \cos(\omega_d t - x_a) \right] \quad (3-17a)$$

$$V_{2m} = \frac{k_2 \beta_2 \rho_2 V_o}{\sqrt{2}} \left[-C_f \sin(\omega_d t + x_f) + C_a \sin(\omega_d t - x_a) \right. \\ \left. + \frac{\gamma \alpha \rho_2 V_o}{k_2} \cos \omega_c t \right] \quad (3-17b)$$

where β_1 and β_2 take care of the conversion losses of the mixers in channels 1 and 2, respectively. The signal voltages as expressed in Equations (3-17a) and (3-17b) are amplified by the series of audio-amplifiers in their respective receiver channels. They also suffer attenuation through the presence of the switches and filters. If the total gains after the 20-dB step amplifiers are G_1 and G_2 , respectively, the variable gain amplifiers are adjusted such that

$$G = \beta_1 k_1 \rho_1 V_O G_1 G_{1v} = \beta_2 k_2 \rho_2 V_O G_2 G_{2v} \quad (3-18)$$

where G_{1v} and G_{2v} are the gains of the variable gain amplifiers in channel 1 and 2, respectively. The signal voltages at the variable gain amplifier output are

$$V_{1v} = \frac{G}{\sqrt{2}} \left[C_f \cos(\omega_d t + x_f) + C_a \cos(\omega_d t - x_a) \right] \quad (3-19a)$$

$$V_{2v} = \frac{G}{\sqrt{2}} \left[-C_f \sin(\omega_d t + x_f) + C_a \sin(\omega_d t - x_a) + \frac{\gamma \alpha \rho_2 V_O}{k_2} \cos \omega_c t \right] \quad (3-19b)$$

The signals as expressed by Equations (3-19a) and (3-19b) are recorded by a tape recorder on two adjacent tracks of the same head. If additional gain unbalance is introduced by the tape recorder, it can be adjusted again in the data reduction. Since only the relative power levels of the backscattered and the calibrate signals are used in the calculations of the backscattering coefficients, the gain adjustment does not affect the final results.

The fore and aft Doppler signals are separated in the process of data reduction. The recorded data are first digitized by an analog-to-digital converter. The Fast Fourier Transform (FFT) is then applied to the digitized data to convert the signals from the time domain to the frequency domain. The frequency spectra of the signals given by Equations (3-19a) and (3-19b) after the application of the FFT are:

$$V_1 = \frac{G}{\sqrt{2}} \left[C_f \{ \delta(f - f_d) \cos x_f - j \delta(f - f_d) \sin x_f \} + C_a \{ \delta(f - f_d) \cos x_a + j \delta(f - f_d) \sin x_a \} \right] \quad (3-20a)$$

$$V_2 = \frac{G}{\sqrt{2}} \left[-C_f \{ \delta(f - f_d) \sin x_f + j \delta(f - f_d) \cos x_f \} + C_a \{ -\delta(f - f_d) \sin x_a + j \delta(f - f_d) \cos x_a \} + \frac{\gamma \alpha \rho_2 V_o'}{k_2} \delta(f - f_c) \right] \quad (3-20b)$$

where $j = \sqrt{-1}$, $f = \frac{\omega}{2\pi}$, and $\delta(f - f_d)$ is the Dirac delta function. In the above expressions, the double-sided spectra (with both positive and negative frequencies) after the FFT are converted to the single-sided spectra by the application of the power scaling factor of 2 (ref. 13). The signal voltage densities, E_f and E_a , (in volts per Hz) at Doppler frequency f_d in the fore and aft direction are obtained by forming the following expressions.

$$E_f = C_f G = \sqrt{\frac{1}{2} \left[\{R_e(V_1) - I_m(V_2)\}^2 + \{R_e(V_2) + I_m(V_1)\}^2 \right]} \quad (3-21a)$$

$$E_a = C_a G = \sqrt{\frac{1}{2} \left[\{R_e(V_1) + I_m(V_2)\}^2 + \{R_e(V_2) - I_m(V_1)\}^2 \right]} \quad (3-21b)$$

The calibrate signal E_c (in volts per Hz) is given by

$$E_c = \frac{G \gamma \rho_2 V_o'}{k_2 \sqrt{2}} \quad (3-22)$$

From Equations (3-21b) and (3-22), the received power density (in watts per Hz) is

$$W_{ra}(f_d) = \frac{C_a^2}{2} = \left(\frac{\gamma \alpha \rho_2 V_o}{2k_2} \right)^2 \left(\frac{E_a}{E_c} \right)^2 \quad (3-23)$$

Since the transmit power P_t is proportional to V_o^2 , Eq. (3-23) can be rewritten as

$$W_{ra}(f_d) = A \left(\frac{\gamma \alpha \rho_2}{2k_2} \right)^2 \left(\frac{E_a}{E_c} \right)^2 P_t = \frac{P_t}{K} \left(\frac{E_a}{E_c} \right)^2 \quad (3-24)$$

where K is a measurable quantity in the laboratory, as discussed in subsection 3.4.

3.3 THE ROLLOFF FUNCTION

From Eq. (3-3), it is seen that the power level P_r of the backscattered signal depends on the backscattering coefficient σ^0 , the distance R from the scatterometer to the surface area of interest, and the gains of the transmit and receive antennae. σ^0 is generally observed to decrease with the angle of incidence θ over either land or water. And $R = h/\cos\theta$, h being the aircraft altitude, increases with θ . Since P_r is proportional to σ^0 and inversely proportional to the fourth power of R , P_r is expected to decrease rapidly with θ . This dependence of P_r on θ has an important bearing on the design and performance of the 1.6 GHz scatterometer system. For example, the

signal to noise ratio of the Minicom 110 Tape Recorder System described in subsection 2.5 is limited to a maximum of 40 dB. If P_r at $\theta = 5^\circ$ is 40 dB or more higher than that at $\theta = 60^\circ$, then the signal power level would be either saturated at $\theta = 5^\circ$ or buried in the noise at $\theta = 60^\circ$ during data recording. Although the antenna array could be designed in such a way that the antenna gains would offset some of the strong dependence of P_r on θ , the approach does not provide enough flexibility for the observations on waters or lands with varying degrees of surface roughness and dielectric property. For the 1.6 GHz scatterometer, the variations of the two-way antenna gains over the incident angles of interest, i.e., from $\theta = 0^\circ$ to $\theta = 60^\circ$, are not more than $\sim 6-7$ dB, as discussed in section 4.

To reduce the power level of the signal return at small θ before being recorded by the tape recorder, two rolloff filters were installed in each of four receiver data channels. One of the filters was used for observations over lands and the other over waters. Either one of the two filters is made operational by a control switch. The responses of these filters as a function of frequency for all four receiver channels were measured in the laboratory at Building 15 of JSC and given in Table A-2 of Appendix A. The frequencies at which the filter responses were measured can be converted to the angles of incidence through Doppler relation of Eq. (3-4) for a given aircraft speed. It is clear from the table that, for a given frequency and switch position, the responses

of the rolloff filters for all four receiver channels are quite similar, the differences being not more than 0.5 dB. For simplicity in later numerical calculations, the average rolloff responses of all four receiver data channels for both land and water filters were derived and plotted in Figure 3-4. Over the frequency range of 35 to 1800 Hz, the attenuation of the receiver output signal varies from ~ 5.6 dB to ~ 0.3 dB for land rolloff filters, while that for the water rolloff filters from ~ 29 dB to ~ 0 dB.

As an example in the application of the rolloff functions $Z_w(\theta)$ and $Z_l(\theta)$ given in Figure 3-4, observe that the values of σ^0 over calm water as obtained from Figure 3-5 in subsection 3.5 are ~ 2 dB and ~ -30 dB for $\theta = 5^\circ$ and $\theta = 60^\circ$, respectively. Assuming that the aircraft is cruising with a speed of ~ 75 m/sec and at an altitude of ~ 480 m, the Doppler frequencies are 70 Hz and 693 Hz, and the distances are ~ 482 m and ~ 960 m for $\theta = 5^\circ$ and $\theta = 60^\circ$, respectively. The corresponding attenuations at these two frequencies are found from Figure 3-4 to be 29 dB and 1.5 dB. If area $A = 1 \text{ m}^2$, $P_t = 1$ watt, and neglecting the signal losses and amplifications in the scatterometer system, the receiver output powers at $\theta = 5^\circ$ and $\theta = 60^\circ$ were calculated from Eq. (3-3) and given in Table 3-1. It is clear that the difference in powers at these two frequencies is only ~ 11.5 dB which certainly does not present a serious problem in data recording. If the rolloff

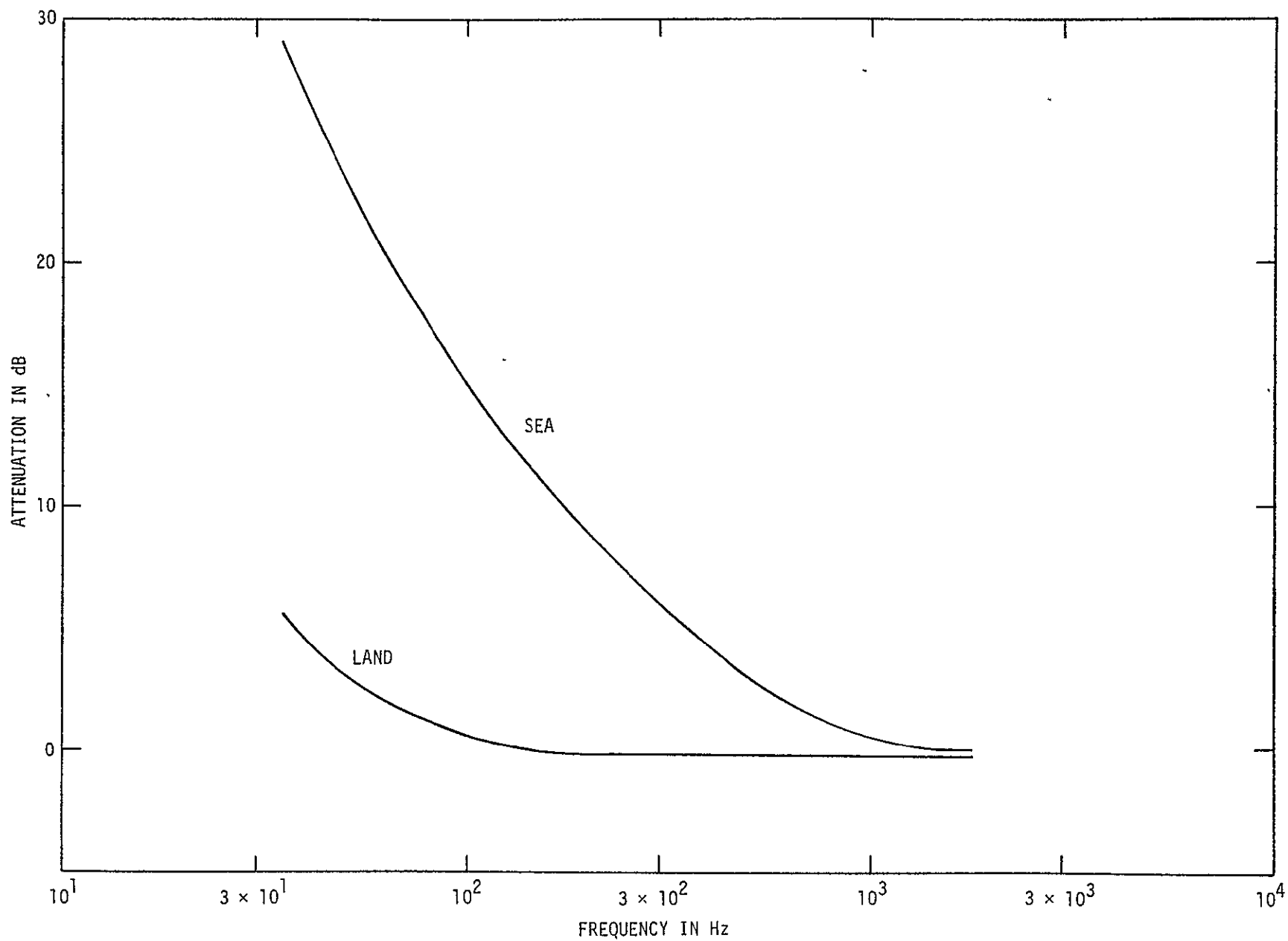


Figure 3-4.— The 1.6 GHz Scatterometer Rolloff Functions.

TABLE 3-1. - A NUMERICAL EXAMPLE OF THE BACKSCATTERED
SIGNAL LEVELS AT THE RECEIVER OUTPUT - HH POLARIZATION

Parameters From Eq. (3-13)	$\theta = 5^\circ$ $f_d = 70 \text{ Hz}$	$\theta = 60^\circ$ $f_d = 693 \text{ Hz}$
P_t	1 Watt, 30 dBm	1 Watt, 30 dBm
$G_t G_r$	19.6 dB	24.6 dB
σ^0	2 dB	-30 dB
λ^2	0.035m^2 , -14.6 dB	0.035m^2 , -14.6 dB
$(4\pi)^3$	33 dB	33 dB
R^4	$5.35 \times 10^{10}\text{m}^4$, 107.3 dB	$8.46 \times 10^{11}\text{m}^4$, 119.3 dB
A	1m^2 , 0 dB	1m^2 , 0 dB
$Z_w(\theta)$	29 dB	1.5 dB
$P_r(\theta)$	-132.3 dBm	-143.8 dBm

filter is not in the system, then the difference in powers would be ~ 39 dB which is undesirably close to the tape recorder dynamic range of ~ 40 dB.

It was found out from many observations over lands that the use of water rolloff filters generally yielded better results for the 1.6 GHz scatterometer. Therefore, it is reasonable to assume that the water rolloff filters will be most frequently used in the future missions either over land or over water. Only the water rolloff function $Z_w(\theta)$ will be used in the numerical calculations in section 5.

3.4 THE MEASURED SYSTEM PARAMETERS

From the results of discussions in the previous three subsections, it is now possible to put all the parameters together and derive the final forms of the data processing equations. Note that both the backscattered and the calibrate signals, E_a and E_c in Equations (3-21B) and (3-22), pass through the water rolloff filters in the receiver data channels and, therefore, subject to the attenuations of the filters amounting to $Z_w(f)$ (or $Z_w(\theta)$). E_c occurs at the frequency of 1.9 KHz so that the value of Z_w is ~ 0 dB. Consequently, only the $Z_w(f)$ for E_a needs to be taken into account and Eq. (3-24) should be rewritten as

$$W_{ra}(f_d) = P_t Z_w(f_d) \left(\frac{E_a}{E_c} \right)^2 \frac{1}{K} \quad (3-25)$$

Combining Equations (3-6), (3-7), and (3-25), the expression for the first form of the data processing equations becomes

$$\frac{Z_{\omega}(f_d)}{K} \left(\frac{E_a}{E_c} \right)^2 \Delta f_d = \frac{\lambda^2}{(4\pi)^3} \sum_{i=1}^N \frac{G_t(\theta_i) G_r(\theta_i) \sigma^0(\theta_i)}{R^4(\theta_i)} \Delta A_i \quad (3-26)$$

Similarly, the second form of the data processing equations is obtained by combining Equations (3-11) and (3-25). The result is

$$\sigma^0(\theta) = \frac{(4\pi)^3 2VhRZ_{\omega}(\theta)}{KG_t(\theta)G_r(\theta)\lambda^3\Delta\phi} \left(\frac{E_a}{E_c} \right)^2 \quad (3-27)$$

Note that, in Eq. (3-26), $E_a^2 \Delta f_d$ is the relative receiver output power in the Doppler frequency band Δf_d corresponding to the signal return from the ground cell defined by two isodopplers, $f_d + \frac{1}{2}\Delta f_d$ and $f_d - \frac{1}{2}\Delta f_d$.

Three key system parameters were measured in Building 15 of JSC, namely, K , $Z_{\omega}(\theta)$ and the cable loss L . The results of the recent measurements of these parameters were given in Appendix A. The rolloff filters were permanently installed in the scatterometer system. Consequently, the values of $Z_{\omega}(\theta)$, which were presented in the previous section, are not expected to change much from measurements at different times unless the filters were replaced. The values of K and L , on the other hand, could change for different missions. The parameter K controls the power level of the

calibrate signal which is also subject to the limitation of the tape recorder dynamic range. For a given surface of backscattering and aircraft altitude, K could be adjusted for an appropriate calibrate signal level to bring about the best signal to noise ratio for the data. The value of L depends on the length of the cables between the R/T unit and the antennae, which could be changed from mission to mission. The values of K and L measured for Mission 347 functional check flight as presented in Appendix A will be used for the numerical calculations in section 5.

Incorporating the cable loss into Equations (3-26) and (3-27), the final forms of the data processing equations become

$$\frac{Z_{\omega}(f_d)}{LK} \left(\frac{E_a}{E_c} \right)^2 \Delta f_d = \frac{\lambda^2}{(4\pi)^3} \sum_{i=1}^N \frac{G_t(\theta_i) G_r(\theta_i) \sigma^0(\theta_i)}{R^4(\theta_i)} \Delta A_i \quad (3-28)$$

and

$$\sigma^0(\theta) = \frac{(4\pi)^3 2VhRZ_{\omega}(\theta)}{LK G_t(\theta) G_r(\theta) \lambda^3 \Delta \phi} \left(\frac{E_a}{E_c} \right)^2 \quad (3-29)$$

3.5 THE FUNCTIONAL DEPENDENCE OF L-BAND σ^0 ON θ

It was pointed out previously that the ordinary approach of deriving the backscattering coefficient σ^0 , by means of a Doppler radar such as the 1.6 GHz scatterometer, is to use the second form of the data processing equation given by Eq. (3-29). The σ^0 derived

from this approach contains two sources of error, both resulting from the approximation of a rectangular area (Figure 3-2) for an actual doppler cell defined by two isodopplers (Figure 3-1).

The first source of error comes from the assignment of a single value of σ^0 for a given rectangular ground cell. Actually, σ^0 was observed to vary with incident angle θ , and the contribution to the observed signal return within a given bandwidth Δf_d is the sum of the backscattered signals from all the small area elements with different θ_i 's in the doppler cell.

The second source of error is due to the crudeness of the area approximation itself. Both of the error sources depend on the scatterometer antenna beamwidth. For the 13.3 GHz scatterometer system, the antenna beamwidth is only $\sim 3^\circ$ and the error in the derivation of σ^0 by Eq. (3-24) is not expected to be large. For the 1.6 GHz scatterometer, however, the antenna beamwidth varies from $\sim 8^\circ$ to $\sim 16^\circ$ over the range of θ from 0° to -60° (see Figure 4-6). In this case, it is desirable to explore the amount of error introduced in the σ^0 derivation by using the second form of data processing equation. The σ^0 derived from the first form of data processing equation given by Eq. (3-28) provide an adequate standard for comparison.

The derivation of σ^0 by Eq. (3-28) requires an assumed functional dependence of σ^0 on θ to begin with, as pointed out previously. To see how the shape of σ^0 versus θ curve affects the error in the

σ^0 estimation by Eq. (3-29), the measured values of σ^0 from both smooth and rough surfaces of backscattering are considered. For the smooth surface, the measurements by Guinard (ref. 14) at the frequency of 1.4 GHz and for VV polarization over calm sea are adopted. For the rough surface, the measurements of Bařlivala and Ulaby (ref. 15), over land with 4.1 cm RMS height and 0.4 g/cm³ soil moisture content in top 1-cm layer, at the frequency of 1.1 GHz and for HH polarization are used. The results from both of these measurements are shown in Figure 3-5. The measurement of σ^0 over the calm sea at $\theta = 86^\circ$ is not included in the figure because of its low value (~ -44 dB). Both of the measurements over land and over water were not made at the frequency of the 1.6 GHz scatterometer. However, this difference in frequencies is not important in the present case where the primary interest is to have the functional forms of σ^0 dependence on θ for the numerical calculations of Eq. (3-28).

A quadratic regression was performed on the data from the measurements over the calm sea. The result is given by

$$\sigma^0 = 7.24 \times 10^{-3} \theta^2 - 1.03\theta + 6.94 \quad (3-30)$$

The data from the measurements over land (ref. 15) covered the range of θ from 0° to 30° only. A simple straight line was drawn through these data points and extended to $\theta \simeq 80^\circ$. This simple-minded approach again is not crucial in the present context for

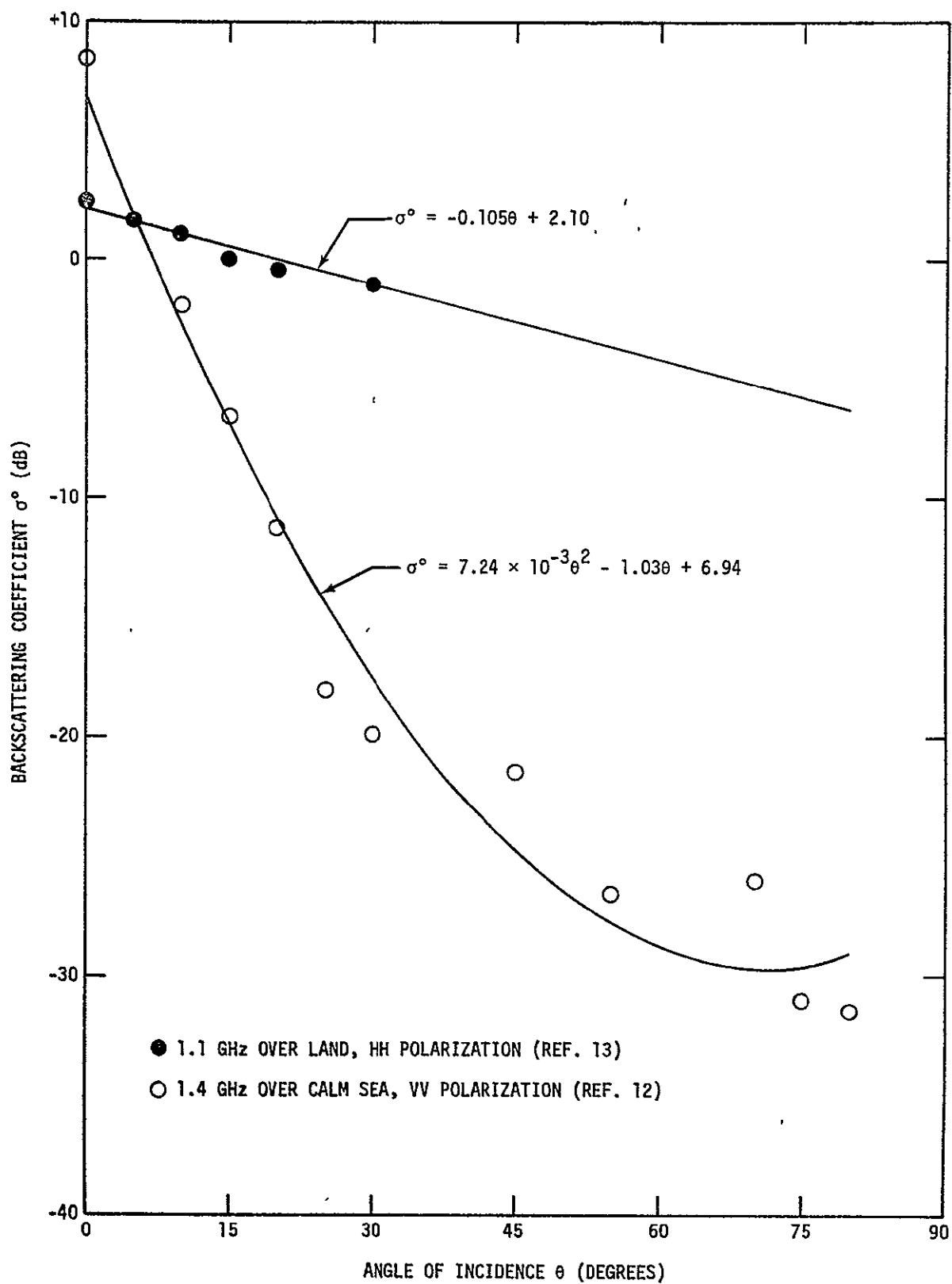


Figure 3-5.— The functional dependence of σ° on θ .

the same reason given in the last sentence of the previous paragraph. The linear dependence of σ^0 on θ over land is expressed by

$$\sigma^0 = -0.105\theta + 2.10 \quad (3-31)$$

In both of the above equations, the unit of σ^0 is in dB and that of θ in degree. Both of these expressions are used in the numerical calculations in subsection 5.2, where the two forms of the data processing equations are compared.

4. THE ANTENNA PATTERNS

4.1 THE MAIN LOBE GAIN

The antenna gain patterns for all four antennae were measured by the Physical Sciences Laboratory of the New Mexico State University for every 2° interval in both along-track and cross-track directions. The measurements were made with a set of four identical antennae without mockup to simulate the actual antenna set mounted on the C130 aircraft. Consequently, the antenna patterns used in the present analysis might deviate from the real ones. The effect resulted from this deviation on the measurements of the backscattering coefficient σ^0 is not known. Calibration of the airborne system against the ground based system may be necessary in order to determine the seriousness of this effect.

To see the spatial variations of the antenna patterns, the contour maps of the one-way gain are displayed in Figures 4.1, 4.2, 4.3, and 4.4, respectively, for the horizontal transmitter, horizontal receiver, vertical transmitter, and vertical receiver. Each of the four maps covers the angular ranges of ϕ from -40° to $+40^\circ$ and of θ from -60° to $+60^\circ$. The direction of nadir is located at $\theta = 0$ and $\phi = 0$. From these figures it is clear that the RF powers of transmission and reception are mostly limited in the regions of ϕ from -20° to $+20^\circ$, although the side lobe level of the vertical receiver antenna appears rather high in the aft direction at $\phi \approx -10^\circ$.

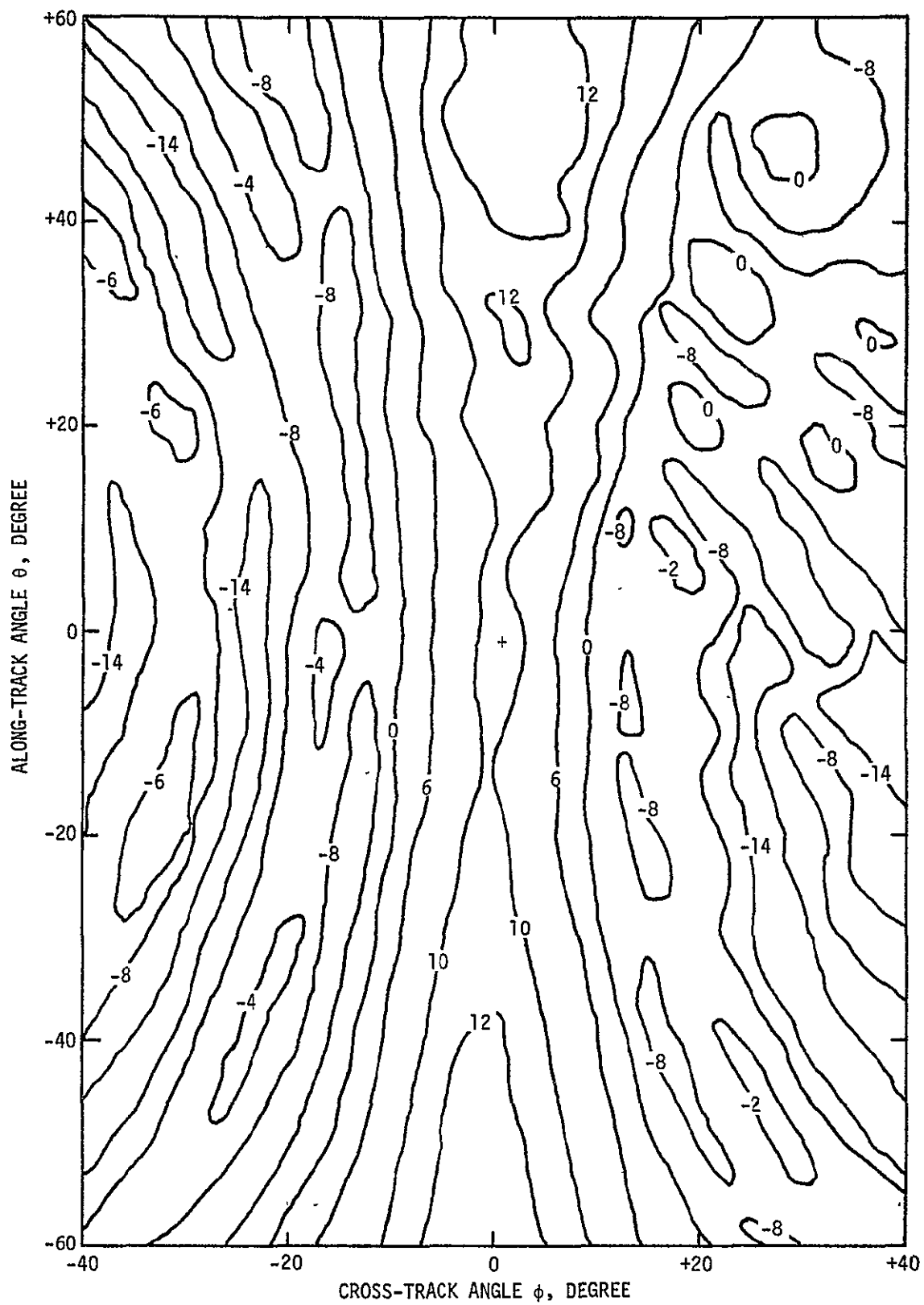


Figure 4-1.— The one-way gain contour of the horizontal transmit antenna.

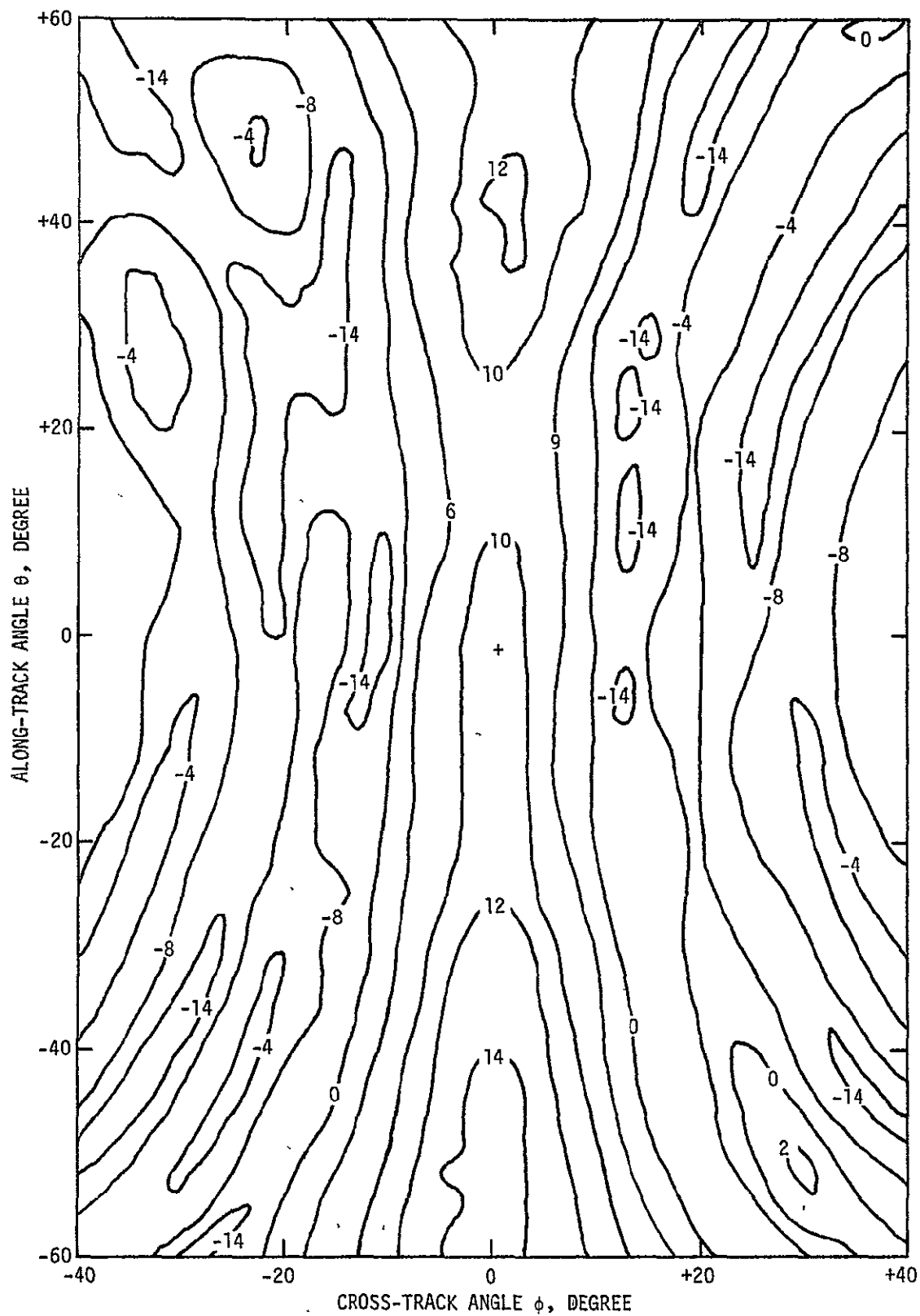


Figure 4-2.—The one-way gain contour of the horizontal receive antenna.

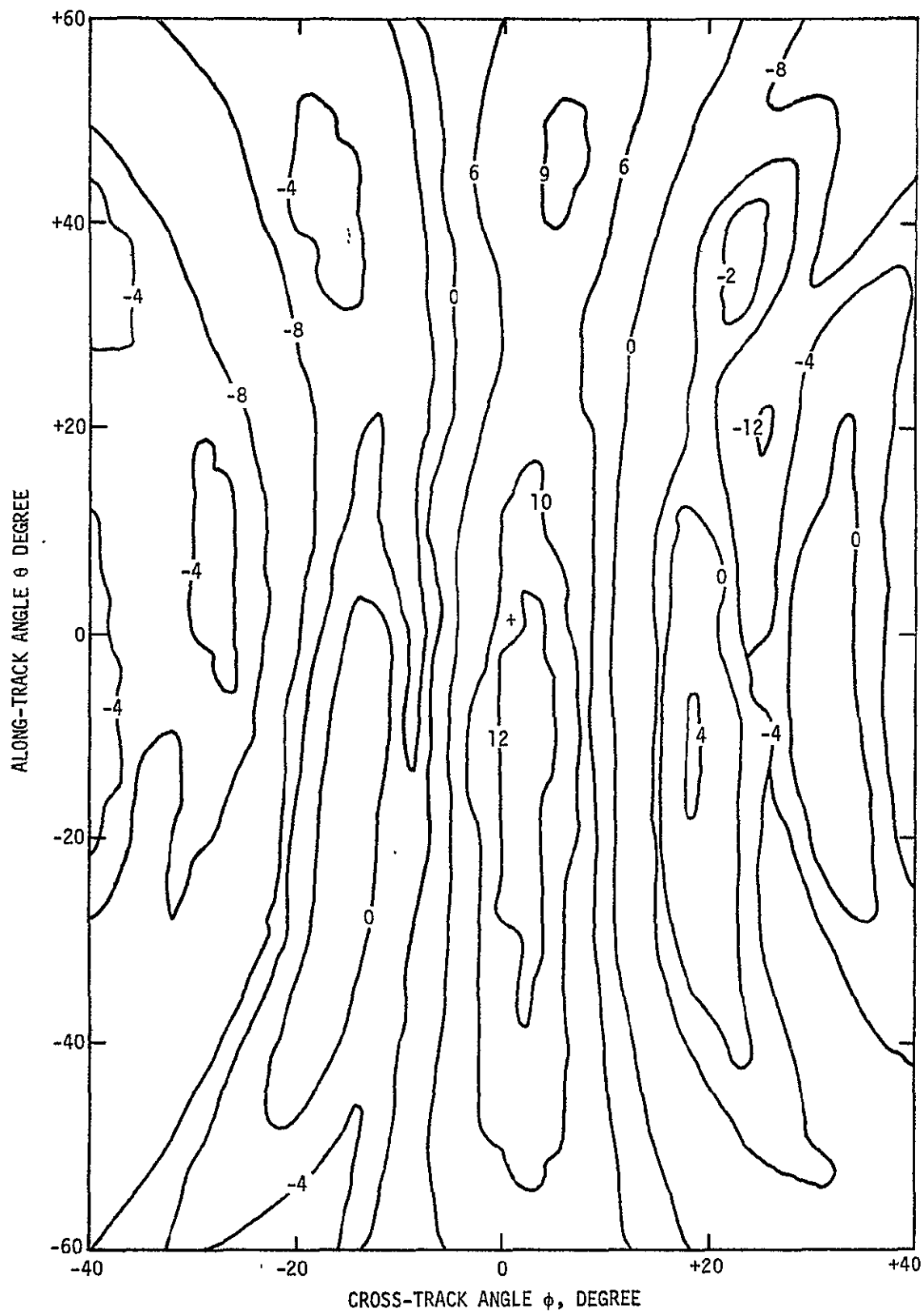


Figure 4-4.— The one-way gain contour of the vertical receive antenna.

The patterns for both horizontal transmitter and receiver antennae shown in Figures 4.1 and 4.2 display an increase in gain with incident angle θ and peaked around $\theta \approx 50^\circ - 60^\circ$ and $\theta \approx -50^\circ - -60^\circ$. On the other hand, the patterns for both vertical transmitter and receiver antennae are quite different from those for their horizontal counterparts. The vertical transmitter antenna gain contour shown in Figure 4.3 clearly indicates the main lobe gain maxima at $\theta \approx -40^\circ$ and $\theta \approx 15^\circ$. The vertical receiver antenna gain contour displayed in Figure 4.4 shows the main lobe gain maxima at $\theta \approx -15^\circ$ and $\theta \approx 45^\circ$. The occurrence of maximum gains at incident angles of $\sim \pm 15^\circ$ might effectively reduce the signal to noise ratio at large incident angles of $\sim \pm 50 - \pm 60^\circ$, as discussed in the following paragraphs.

The variations of the along-track two-way gains with the angles of incidence θ in the aft direction for all four combinations of polarization states are shown in Figure 4.5. It is clear from the figure that the maxima of the two-way gains for HH (horizontal transmit and horizontal receive) and VH (vertical transmit and horizontal receive) combinations appear at $\theta \approx -55^\circ$ and $\theta \approx -45^\circ$, respectively. On the other hand, the gain maxima for both HV (horizontal transmit and vertical receive) and VV (vertical transmit and vertical receive) combinations are found to be at $\theta \approx -40^\circ$. This occurrence of gain maxima at an incident angle of $\sim -40^\circ$ is due to the gain pattern of the vertical receive antenna shown in Figure 4.4. From Eq. (3-3) of the previous section, it

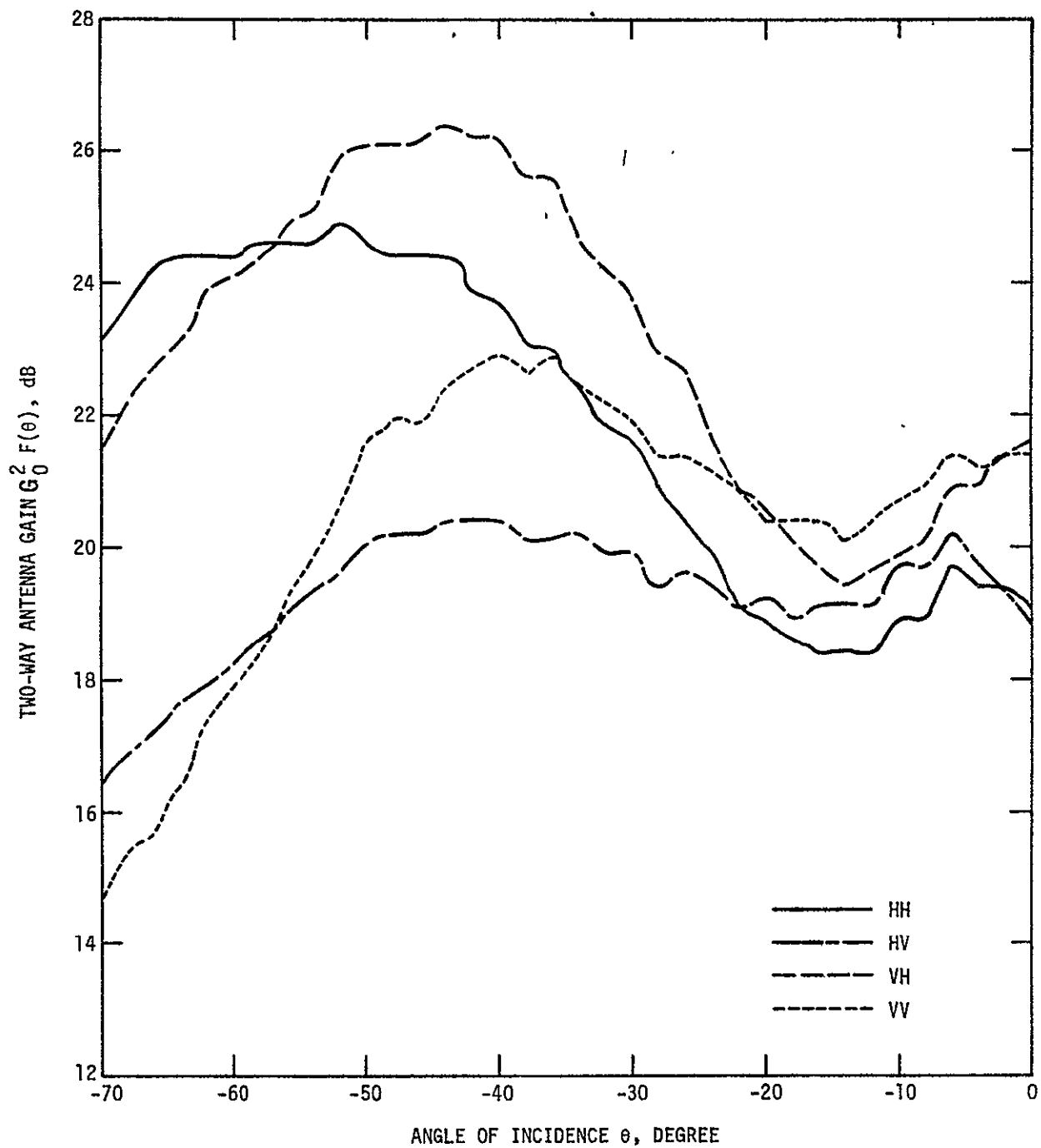


Figure 4-5.— The variations of two-way antenna gains with angle of incidence.

is observed that the backscattered power P_r is proportional to $G_t G_r$, σ^0 and R^{-4} . Since both σ^0 and R^{-4} generally decrease with increasing $|\theta|$ (absolute value), P_r would be small at large $|\theta|$. For both HV and VV polarization combinations, the decrease in two-way antenna gain $G_t G_r$ for $|\theta| > 40^\circ$ further depresses the level of P_r . Thus, for those two combinations, the signal to noise ratio is expected to decrease with increasing $|\theta|$ for $|\theta| > 40^\circ$.

The beamwidth variations of the two-way main lobe antenna gains were also examined here. There were many ways to define the antenna beamwidth, some of them were discussed by Krishen et al. (ref. 4). In the following discussion, two methods of defining the antenna beamwidth were adopted. The first one is to take the maximum value of the two-way gain $G_o^2 F(\theta)$ at an incidence angle θ . A 3-dB value was subtracted from $G_o^2 F(\theta)$ (in dB). Two values of the cross-track angle ϕ_1 and ϕ_2 , measured from the location of $G_o^2 F(\theta)$, were determined where the antenna gain equaled $G_o^2 F(\theta) - 3$ dB. The 3-dB beamwidth is the sum of ϕ_1 and ϕ_2 . The results from this method are shown in Figure 4.6 for all four combinations of polarizations. Two features are immediately clear from this figure. First, the beamwidth variations with θ are similar for all four combinations. For a given θ , the difference in beamwidth for any two combinations is in the order of $\sim 1^\circ - 2^\circ$ or less. Secondly, for small incident angles, i.e., $|\theta| < 30^\circ$, the 3-dB beamwidth is $\sim 8^\circ \pm 1^\circ$ and remains

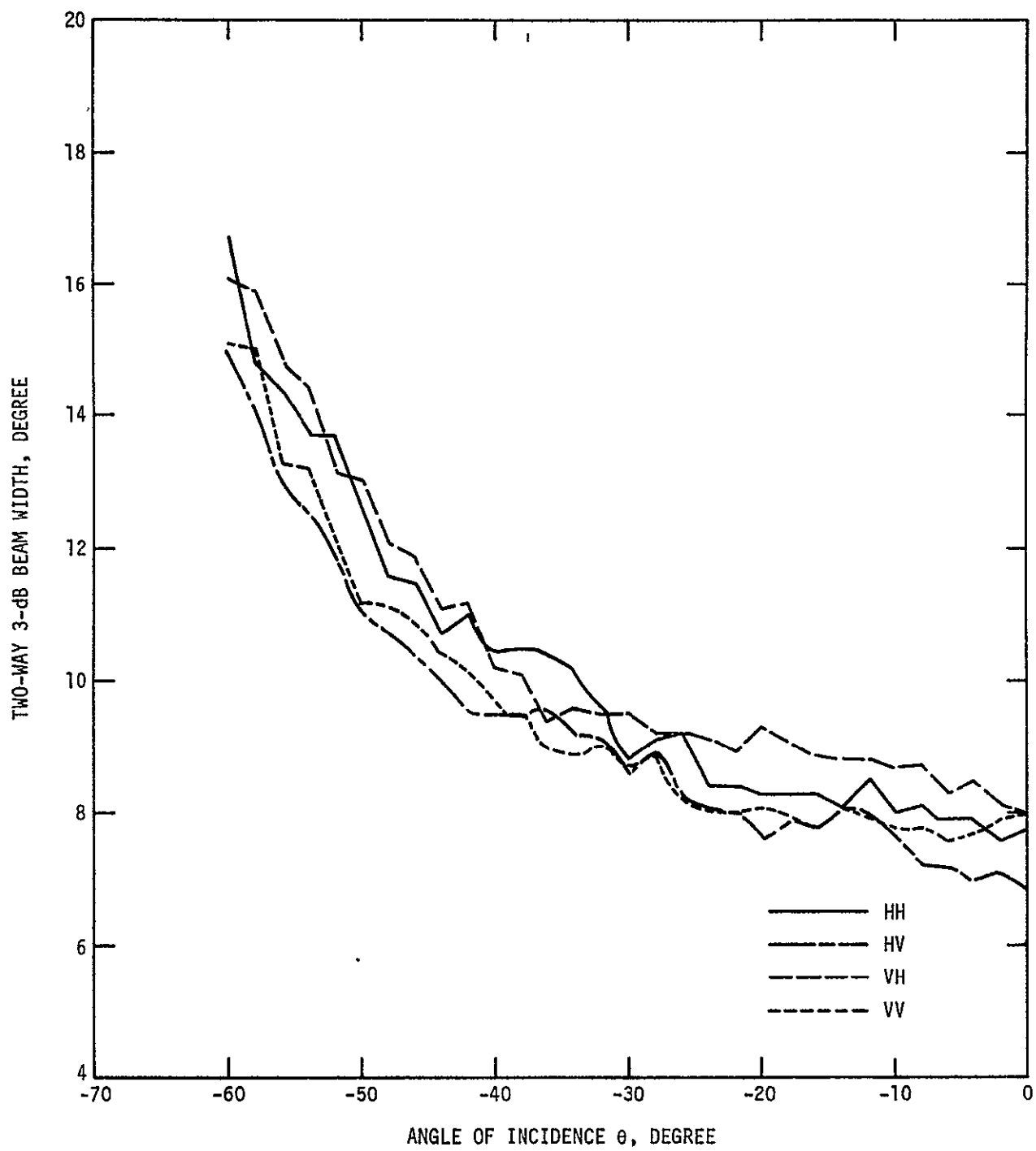


Figure 4-6.— The variations of the 3-dB beam width with angle of incidence.

relatively constant with θ . For $|\theta| > 30^\circ$, the beamwidth increases with incident angle rapidly. The value of the beamwidth varies from $\sim 9^\circ$ at $\theta \sim -30^\circ$ to $\sim 16^\circ$ at $\theta \sim -60^\circ$.

The second method of defining a beamwidth is given by

$$\Delta\phi = \frac{\int_{\phi_1}^{\phi_2} G_o^2 F(\theta) f(\theta, \phi) d\phi}{G_o^2 F(\theta)} \quad (4-1)$$

where ϕ_1 and ϕ_2 are the cross-track angles where the null points of the main lobe antenna gain occur. The function $f(\theta, \phi)$ defines the cross-track gain variation and $G_o^2 F(\theta)$, which is expressed in numerical value and not in dB, has the same meaning as before. The results of the calculations based on Eq. (4-1) are tabulated in Table 4.1 for incident angles ranging from 0° to -60° at every 2° interval. The corresponding 3-dB beamwidths derived in the previous paragraph are also included in the table for comparison. It is clear from the table that, for all angles of incidence and all polarization combinations, the values of the beamwidth as derived from Eq. (4-1) are larger than the corresponding ones obtained from the 3-dB method. On the average, the difference in the beamwidth calculated by the two methods is $\sim 1^\circ$.

The antenna beamwidth as defined by Eq. (4-1) will be used in the numerical calculations discussed in Section 5.

TABLE 4-1. - COMPUTED TWO-WAY ANTENNA BEAMWIDTH IN DEGREE

Incident Angle θ (Degree)	Method 1				Method 2			
	HH	HV	VH	VV	HH	HV	VH	VV
0	7.7	6.9	8.0	8.0	8.6	9.8	8.8	9.1
-2	7.6	7.1	8.1	7.9	8.5	9.4	8.8	9.1
-4	7.9	7.0	8.5	7.7	8.7	9.3	9.1	9.0
-6	7.9	7.2	8.3	7.6	8.3	8.5	8.5	8.7
-8	8.1	7.2	8.7	7.8	9.1	9.1	9.3	8.9
-10	8.0	7.6	8.7	7.8	8.6	8.8	9.1	8.5
-12	8.5	8.0	8.8	8.0	9.0	9.3	9.3	8.9
-14	8.1	8.1	8.8	8.1	8.7	9.1	9.2	8.7
-16	8.3	7.8	8.9	7.8	8.9	8.9	9.9	8.7
-18	8.3	7.9	9.1	8.0	8.9	8.8	9.4	8.8
-20	8.3	7.6	9.3	8.1	9.1	8.6	10.0	9.0
-22	8.4	8.0	8.9	8.0	9.0	9.0	9.4	8.7
-24	8.4	8.1	9.1	8.0	9.2	9.3	9.6	8.7
-26	9.2	8.2	9.2	8.2	9.6	9.4	9.9	8.8
-28	9.1	8.9	9.2	8.8	9.9	10.1	10.0	9.5
-30	8.8	8.6	9.5	8.7	9.5	9.8	10.3	9.5
-32	9.6	9.1	9.5	9.0	10.2	10.1	10.3	9.6
-34	10.1	9.2	9.6	8.9	10.1	9.9	10.5	9.6
-36	10.4	9.6	9.4	9.0	11.0	10.5	10.2	9.6
-38	10.5	9.5	10.1	9.5	11.1	10.8	10.9	10.0
-40	10.5	9.5	10.2	9.6	11.1	10.4	11.0	10.1
-42	11.0	9.5	11.2	10.1	11.7	10.8	11.8	10.8
-44	10.7	10.0	11.1	10.4	11.6	11.1	11.9	11.0
-46	11.5	10.4	11.0	10.9	12.6	11.7	12.8	11.5
-48	11.6	10.7	12.1	11.1	12.9	11.9	13.1	11.7
-50	12.6	11.1	13.0	11.2	13.4	12.2	13.7	11.7
-52	13.7	11.9	13.2	12.1	13.8	13.3	13.9	12.9
-54	13.7	12.6	14.4	13.2	14.4	13.6	15.3	13.9
-56	14.4	13.0	14.8	13.3	15.4	14.4	15.8	14.1
-58	14.8	14.1	15.9	15.0	15.7	15.2	17.4	16.1
-60	16.6	15.0	16.1	15.1	17.6	16.3	17.6	16.0

4.2 THE SIDE LOBES

From Figures 4-1, 4-2, 4-3, and 4-4 presented in the subsection above, it appears that the contribution of the power return from the side lobes is not significant compared to that from the main lobe. To examine the relative levels of the main and side lobes more carefully, the two-way gains at an incident angle of -20° are plotted as a function of the cross-track angle ϕ in Figures 4-7 and 4-8 for all four combinations of antenna polarizations. For the HH and HV combinations shown in Figure 4-7, the first side lobe gains are, respectively, ~ 30 dB and ~ 25 dB lower compared to the main lobe gain. In Figure 4-8, the same comparison indicated the differences of ~ 20 dB and ~ 25 dB between the main and the first side lobe gains for the VV and VH combinations, respectively. The gain of each of the remaining side lobes is comparable to that of the first side lobe in each of the four combinations. In general, the side lobe gains become progressively smaller as $|\phi|$ increases. Although the contribution to the power return from each individual side lobe is small, it is desirable to obtain an estimate on the total contribution from all of the side lobes.

The derivation of the backscattering coefficient σ^0 normally takes into account the ground cell associated only with the main lobe antenna pattern. The power return at the scatterometer receiver, however, includes contributions from other areas where the side lobe gains are finite and from the antenna coupling effect discussed in the next subsection. At a given Doppler frequency band

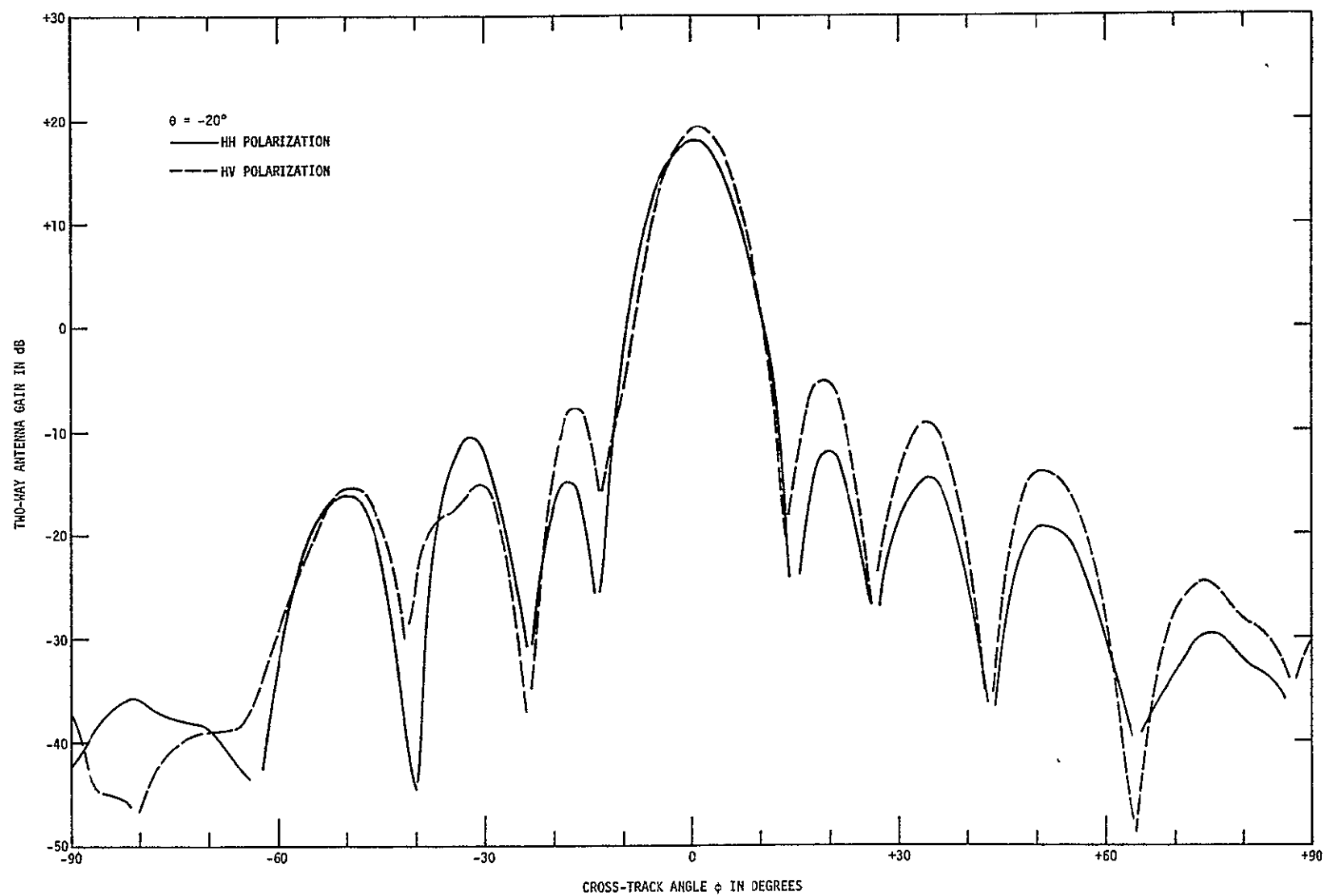


Figure 4-7.— The variation of the two-way antenna gains with cross-track angles.

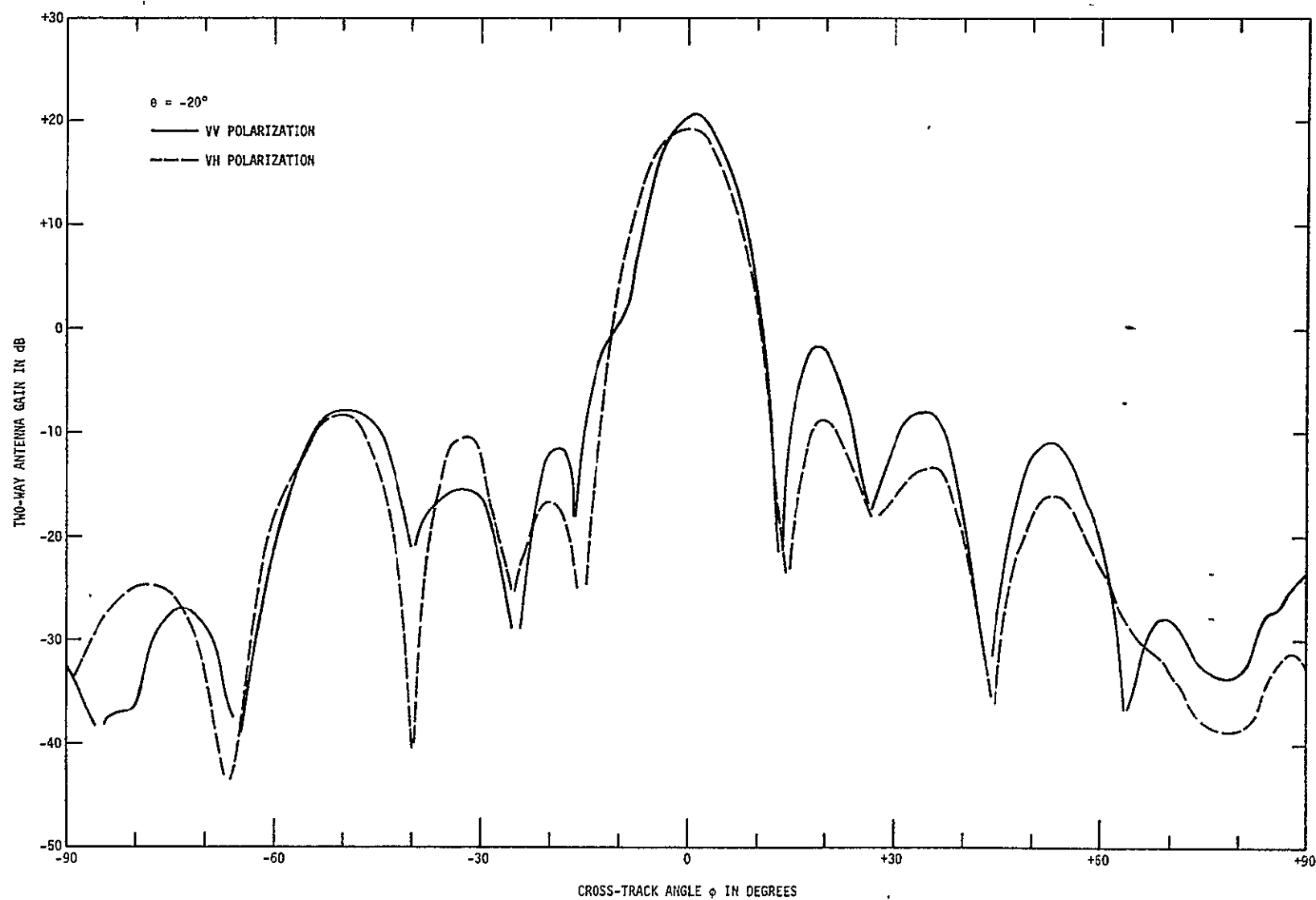


Figure 4-8.— The variations of the two-way antenna gains with cross-track angles.

associated with along-track incident angle θ and angular width $\Delta\theta$, the side lobe contribution to the power return is

$$P_S(\theta) = \frac{P_t \lambda^2}{(4\pi)^3} \int_{A_S} \frac{G_t G_r \sigma^0}{R^4} dA \quad (4-2)$$

where the integration is carried out over the area defined by the Doppler frequency band and all side lobes. To make a rough but reasonable numerical estimate of $P_S(\theta)$, the area element of integration is to be approximated by the one shown in Figure 4-9. The two isodopplers determine the total ground area which back-scatters the RF power in the corresponding Doppler frequency band. Curves L_1 and L_2 represent the surface loci of the 3-dB points of a given side lobe. Point C_i is the intersection between the isodoppler corresponding to the central Doppler frequency and the curve tracing the two-way gain maxima of the side lobe. The area element from this side lobe is approximately given by

$$\Delta A_i = \Delta x_i \Delta y_i \quad (4-3)$$

and Eq. (4-2) can be written as

$$P_S(\theta) = \frac{P_t \lambda^2}{(4\pi)^3} \sum_{i=1}^N \frac{G_i^2 \sigma_i^0}{R_i^4} \Delta x_i \Delta y_i \quad (4-4)$$

where G_i^2 is the side lobe maximum two-way gain at point C_i for the area element ΔA_i . The values of σ_i^0 and R_i all refer to

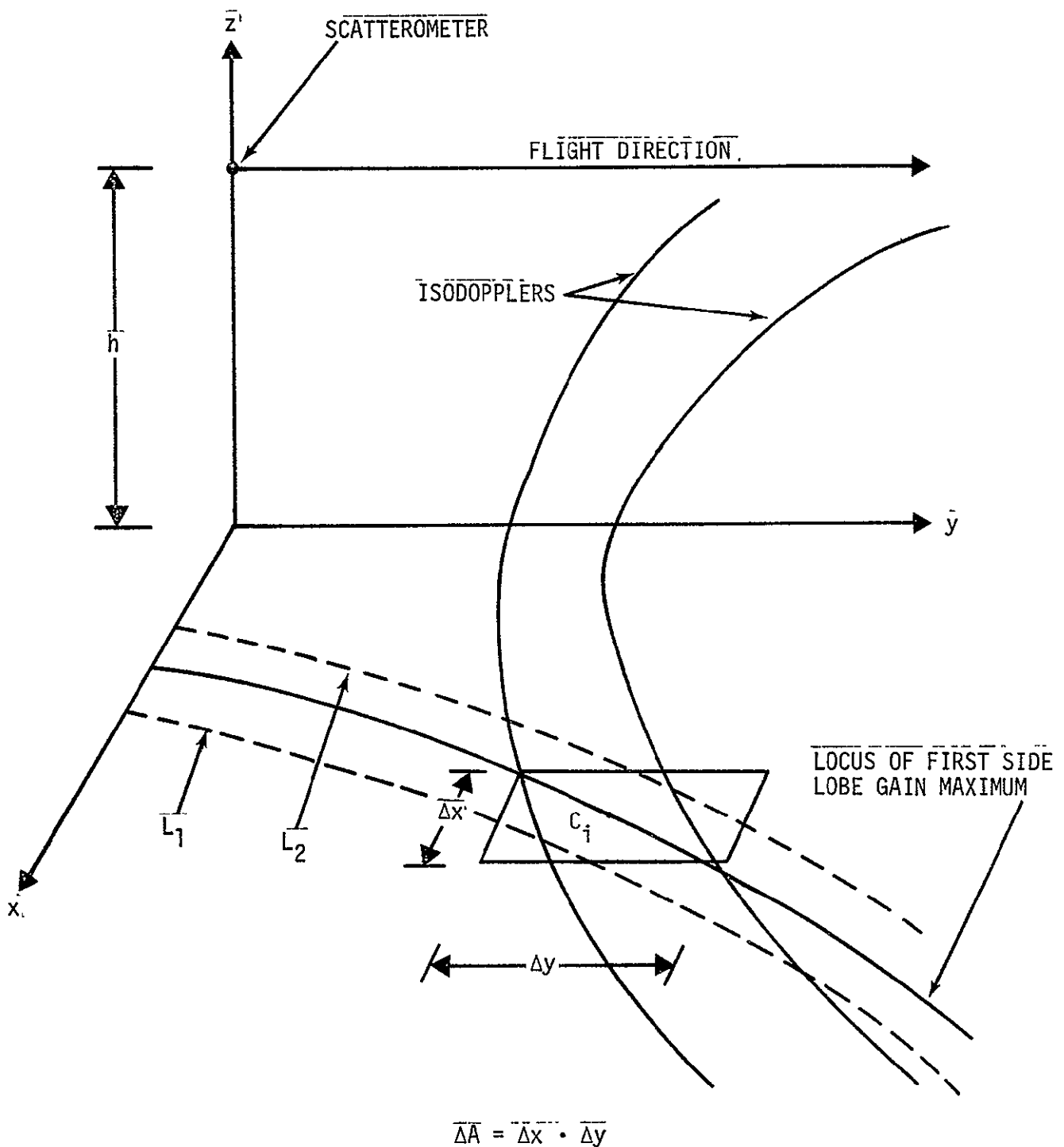


Figure 4-9.— An area element used in the calculation of side lobe contribution to total signal.

the point C_i . The summation is over all the side lobe area elements within the two isodopplers.

The estimated values for $P_s(\theta)$ are presented in Section 5 where results from all numerical calculations are summarized and discussed.

4.3 THE ANTENNA COUPLING

Since the isolation among the two transmit and the two receive antennae is not perfect, undesired signals are generated due to the antenna coupling which appear in the receiver data channels as portions of the backscattered signals. The relative contribution of the undesired signals to the total signal power level has important bearing in the data analysis and the interpretation of the backscattering process of the RF waves. Thus, it is necessary to estimate the relative signal contribution due to the effect of the antenna coupling in order to correctly interpret the observational results.

There are two distinct sources of these undesired signals in the 1.6 GHz scatterometer. The first source comes from the direct power leakage into the receiver data channels due to the coupling between the transmit and receive antennae. The level of this leakage signals is ~ 65 dB below the transmitter power and does not introduce much error in the value of P_t in the data processing equation. Furthermore, the signals from this source are not Doppler-frequency

shifted and would appear as power return in the nadir direction. The signals at zero Doppler frequency shift are generally not used in the derivation of the backscattering coefficient σ^0 from this scatterometer. Therefore, this source of undesired signals will not be discussed in detail here.

The second source of the undesired signals results from the cross polarization effects of each of the two transmit antennae as well as of each of the two receive antennae. The signals from this source are Doppler frequency shifted and are entangled with the genuine backscattered signals of interest. To see how this source of undesired signals are generated, a specific example is described below. Note that the expression for $P_{rVH}(\theta)$, the power level of the signal return at an incident angle θ in the vertical-transmit and horizontal-receive data channel, including the antenna coupling effect, can be written as

$$P_{rVH}(\theta) = \frac{\lambda^2 \Delta A P_t}{(4\pi)^3 R^4(\theta)} \left[G_{tVM}(\theta) G_{rHM}(\theta) \sigma_{VH}^0(\theta) + G_{tVC}(\theta) G_{rHM}(\theta) \sigma_{HH}^0(\theta) + G_{tVM}(\theta) G_{rHC}(\theta) \sigma_{VV}^0(\theta) + G_{tVC}(\theta) G_{rHC}(\theta) \sigma_{HV}^0(\theta) \right] \quad (4-5)$$

where G_{tVM} = vertically-polarized transmit antenna gain when the vertical transmit antenna is on.

G_{tVC} = cross-polarized vertical transmit antenna gain when the vertical transmit antenna is on.

G_{rHM} = horizontally-polarized receive antenna gain when the horizontal receive antenna is on.

G_{rHC} = cross-polarized horizontal receive antenna gain when the horizontal receive antenna is on.

The first term inside the bracket in Eq. (4-5) is the dominant term which represents the backscattered signals of interest. The three remaining terms are contributions of the undesired signals due to antenna coupling. These three terms, in sequential order, represent:

- The return signal in the horizontally-polarized receiver due to the signal transmitted by the cross-polarized component of vertical transmit antenna and backscattered in the horizontal polarization.
- The return signal in the cross-polarized horizontal receiver due to the signal transmitted by the vertically-polarized transmit antenna and backscattered in the vertical polarization.
- The return signal in cross-polarized horizontal receiver due to the signal transmitted by cross-polarized component of vertical transmit antenna and backscattered in the vertical polarization.

Expressions similar to Eq. (4-5) were derived for the other three power returns of different polarization states, namely P_{rHH} , P_{rHV} , and P_{rVV} . The results were summarized in Table 4-2, including the one for P_{rVH} discussed above. For each of the four power returns in the table, the first term always gives the dominant term of interest from which the associated backscattering coefficient σ^0 is derived. The other three terms representing the undesired signal returns would have to be estimated and their relative contributions to the total power level assessed in

TABLE 4-2. - THE COMPONENTS OF THE BACKSCATTERED
SIGNAL IN RECEIVER DATA CHANNELS

Power Return	Components of Backscattered Signal
P_{rHH}	1. $G_{tHH} \sigma^0_{HH} G_{rHM}$ 2. $G_{tHC} \sigma^0_{VH} G_{rHM}$ 3. $G_{tHM} \sigma^0_{HV} G_{rHC}$ 4. $G_{tHC} \sigma^0_{VV} G_{rHC}$
P_{rHV}	1. $G_{tHM} \sigma^0_{HV} G_{rVM}$ 2. $G_{tHC} \sigma^0_{VV} G_{rVM}$ 3. $G_{tHM} \sigma^0_{HH} G_{rVC}$ 4. $G_{tHC} \sigma^0_{VH} G_{rVC}$
P_{rVH}	1. $G_{tVM} \sigma^0_{VH} G_{rHM}$ 2. $G_{tVC} \sigma^0_{HH} G_{rHM}$ 3. $G_{tVM} \sigma^0_{VV} G_{rHC}$ 4. $G_{tVC} \sigma^0_{HV} G_{rHC}$
P_{rVV}	1. $G_{tVM} \sigma^0_{VV} G_{rVM}$ 2. $G_{tVC} \sigma^0_{HV} G_{rVM}$ 3. $G_{tVM} \sigma^0_{VH} G_{rVC}$ 4. $G_{tVC} \sigma^0_{HH} G_{rVC}$

the derivation of σ^0 . In Eq. (4-5) as well as all the calculations in the following, the four backscattered signals are assumed to be incoherent.

To show the level of antenna coupling, the one-way gains of both prime and cross (coupling) polarizations at $\phi = 0^\circ$ for all four antennae were plotted as a function of θ in the aft direction in Figures 4-10 and 4-11. For the horizontal transmit and receive antenna gains shown in Figure 4-10, the cross polarization level is generally about 20 dB or more below that of the prime polarization for all θ . The same assessment can be made for the gains of the vertical transmit and receive antennae shown in Figure 4-11. The average antenna coupling effect is generally stronger at locations where $\phi \neq 0^\circ$. This is illustrated in Figures 4-12 and 4-13 where the one-way gains of both prime and cross polarizations were plotted as a function of ϕ at $\theta = -10^\circ$ in the aft direction for all four antennae. Although the cross-polarization gains are observed to vary with ϕ , the average gain values over ϕ in the range from -20° to $+20^\circ$ are not very different from that at $\phi = 0^\circ$. The gains for the prime polarization, on the other hand, decrease steadily on both sides of $\phi = 0^\circ$.

Based on Figures 4-12 and 4-13 and Table 4-2, the relative contributions due to antenna coupling could be estimated, at some selected locations, for each of the four power returns, P_{rHH} ,

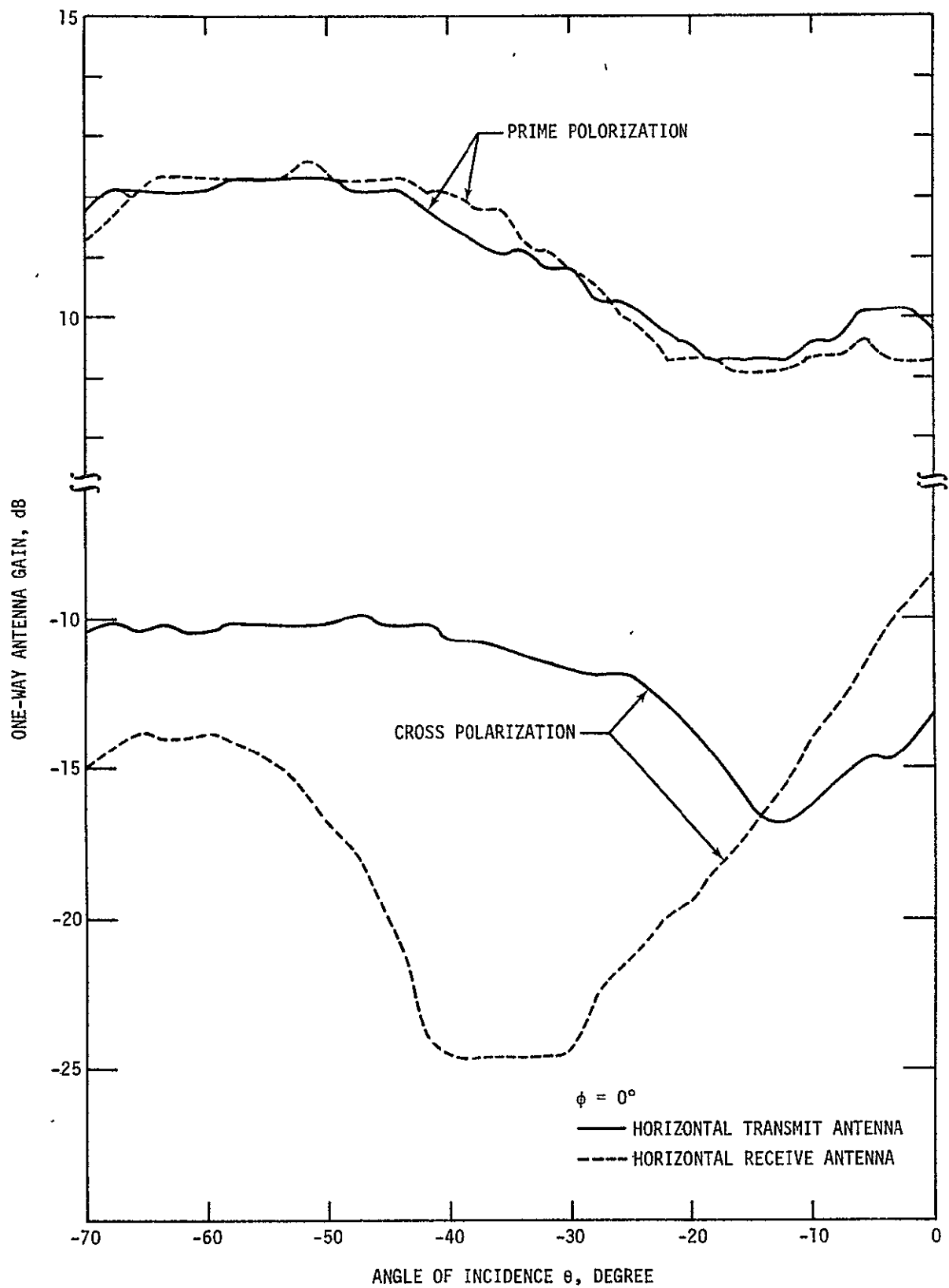


Figure 4-10.— The variations of one-way antenna gain with incident angle θ for both prime and cross polarizations.

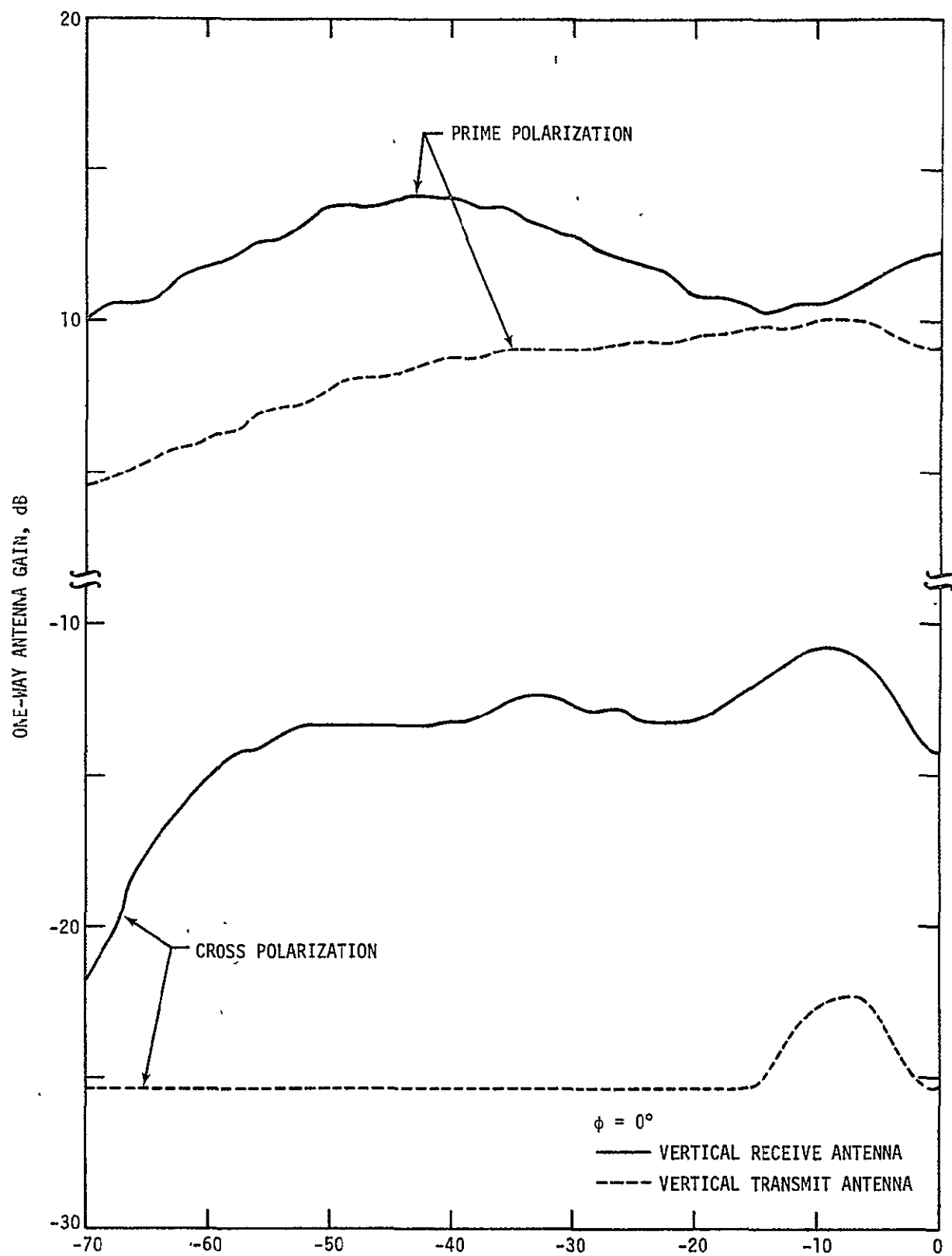


Figure 4-11.— The variation of one-way antenna gain with incident angle θ for both prime and cross polarizations.

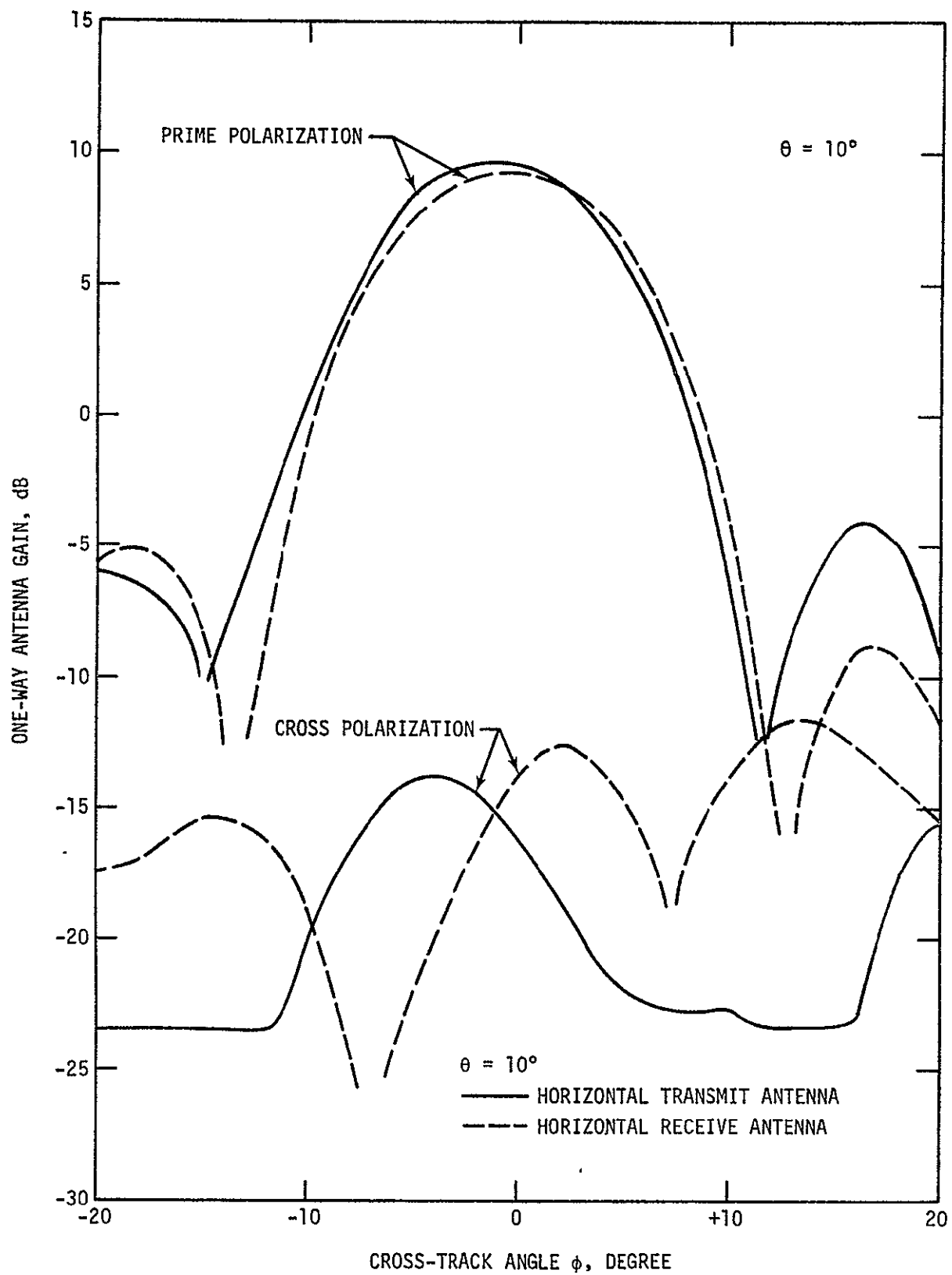


Figure 4-12.— The variation of antenna coupling effect with cross-track angle.

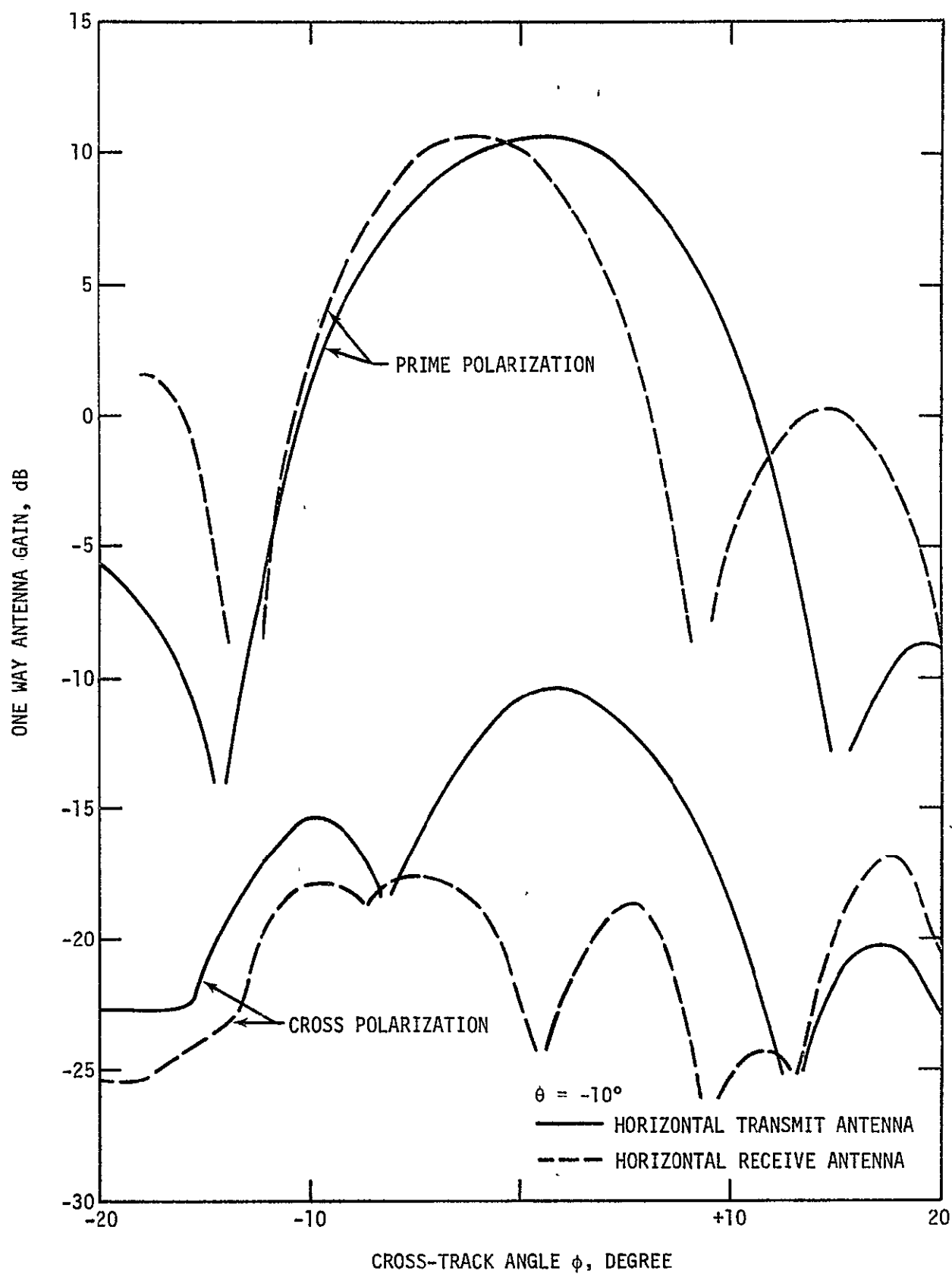


Figure 4-13.— The variation of antenna coupling effect with cross-track angle.

P_{rVV} , P_{rVH} , P_{rHV} . The results are summarized in Table 4-3 for $\theta = -10^\circ$ and for two ϕ values of 0° and 4° (~3 -dB point). All of the relative power returns and σ^0 's in the table are expressed in dB. The components of the backscattered signals are listed in the same sequence as those in Table 4-2. Since the values of σ_{HH}^0 and σ_{VV}^0 are generally observed to be about 10-15 dB greater than those for σ_{VH}^0 or σ_{HV}^0 (ref. 20), it is immediately clear from the table that, for P_{rHH} and P_{rVV} , the contributions from the undesired signals are ~30 dB or more below that from the signal of interest. For the remaining two power returns P_{rHV} and P_{rVH} , the undesired signal contributions may be appreciable. For example, at $\phi = 0^\circ$ and $\phi = -10^\circ$, the power level of the third undesired signal component could be only ~4-9 dB below that of the signal of interest in P_{rHV} . These results suggest that, depending upon locations, the contributions of the undesired signal could be as high as 40% of the total power return for either P_{rHV} or P_{rVH} .

The power at a given Doppler frequency band as observed by the 1.6 GHz scatterometer represents the total backscattered signal from the entire ground cell corresponding to that frequency band. Therefore, the comparisons made in Table 4-3 at certain selected locations are clearly not sufficient to give a valuable assessment of the antenna coupling effect. A reasonable approach to make an estimate of the effect is to integrate the power of each of the backscattered signal components, desired and undesired, over the entire ground cell and compare the results. This is done by numerical integration described in Section 5.

TABLE 4-3. - THE RELATIVE CONTRIBUTIONS TO THE
TOTAL BACKSCATTERED SIGNAL

POWER RETURN	COMPONENTS OF BACKSCATTERED SIGNAL	
	$\phi = 0^\circ, \theta = -10^\circ$	$\phi = 4^\circ, \theta = -10^\circ$
ΔP_{rHH}	1. $\sigma^{HH} + 18.9 \text{ dB}$ 2. $\sigma^{VH} - 6.9 \text{ dB}$ 3. $\sigma^{HV} - 13.1 \text{ dB}$ 4. $\sigma^{VV} - 38.9 \text{ dB}$	1. $\sigma^{HH} + 14.7 \text{ dB}$ 2. $\sigma^{VH} - 13.6 \text{ dB}$ 3. $\sigma^{HV} - 12.3 \text{ dB}$ 4. $\sigma^{VV} - 39.6 \text{ dB}$
ΔP_{rHV}	1. $\sigma^{HV} + 19.7 \text{ dB}$ 2. $\sigma^{VV} - 6.1 \text{ dB}$ 3. $\sigma^{HH} - 4.3 \text{ dB}$ 4. $\sigma^{VH} - 30.1 \text{ dB}$	1. $\sigma^{HV} + 12.4 \text{ dB}$ 2. $\sigma^{VV} - 14.9 \text{ dB}$ 3. $\sigma^{HH} - 6.5 \text{ dB}$ 4. $\sigma^{VH} - 33.8 \text{ dB}$
ΔP_{rVH}	1. $\sigma^{VH} + 19.9 \text{ dB}$ 2. $\sigma^{HH} - 1.4 \text{ dB}$ 3. $\sigma^{VV} - 12.1 \text{ dB}$ 4. $\sigma^{HV} - 33.4 \text{ dB}$	1. $\sigma^{VH} + 17.7 \text{ dB}$ 2. $\sigma^{HH} - 3.6 \text{ dB}$ 3. $\sigma^{VV} - 9.3 \text{ dB}$ 4. $\sigma^{HV} - 30.6 \text{ dB}$
ΔP_{rVV}	1. $\sigma^{VV} + 20.7 \text{ dB}$ 2. $\sigma^{HV} - 0.6 \text{ dB}$ 3. $\sigma^{VH} - 3.3 \text{ dB}$ 4. $\sigma^{HH} - 24.6 \text{ dB}$	1. $\sigma^{VV} + 15.4 \text{ dB}$ 2. $\sigma^{HV} - 5.9 \text{ dB}$ 3. $\sigma^{VH} - 3.5 \text{ dB}$ 4. $\sigma^{HH} - 24.8 \text{ dB}$

It is noted from Figure 4-11 that the measurements of the antenna gains are limited to the levels above ~ -25 dB. This suggests that the actual cross-polarized gain levels could be lower than the measured values in some regions. However, the estimate on the total undesired signal contribution is heavily weighted towards the components having the higher power returns associated with higher antenna cross-polarized levels. Consequently, the saturation in the cross-polarized gain measurements does not introduce significant errors in the numerical estimates of the antenna coupling effect performed in Section 5.

The coupling in the side lobes also makes undesired contributions to the total signal power return. However, the contribution to the total signal due to the prime side lobe gains is shown to be small in Sections 4.2 and 5.3. The undesired signal level due to the antenna coupling in the side lobes is expected to be even smaller and will, therefore, not be explored.

5. NUMERICAL RESULTS AND DISCUSSION

5.1 THE COMPUTER PROGRAM

The numerical calculations of the receiver signal outputs for various polarization states were performed according to Eq. (3-28). The area elements used in the numerical integration covered 2° in ϕ . The center of the area elements lie on the isodoppler corresponding to the along-track incident angle θ_0 . The area element ΔA_i is given by

$$\Delta A_i = \Delta x_i \Delta y_i \quad (5-1)$$

where Δy_i for off-track ($\phi \neq 0$) ΔA_i is related to Δy_0 ($\phi = 0$) along the y-axis by

$$\Delta y_i = \Delta y_0 (1 + \tan^2 \phi_i)^{\frac{1}{2}} \quad (5-2)$$

Δy_0 is taken to be 50 m in the computation. The width Δx_i is defined to be the distance subtended by 2° of the cross-track angular increment and can be written as

$$\Delta x_i = h \left[\tan (\phi_i + 1) - \tan (\phi_i - 1) \right] \quad (5-3)$$

Here h is the aircraft altitude and is taken to be 460 m (1510 feet). The antenna gains $G_t(\theta_i)$ and $G_r(\theta_i)$, the backscattering coefficient $\sigma^0(\theta_i)$ and the distance $R(\theta_i)$ inside the summation

sign of Eq. (3-28) were all evaluated at the centers of ΔA_i 's. When these centers did not fall on the positions where the actual measurements of $G_t(\theta_i)$ and $G_r(\theta_i)$ were made, linearly interpolated values from the immediate neighbors were used. The summation was carried out to the null points on both sides of the main beam.

It was pointed out in Section 4 that, because of antenna coupling, there were undesirable contributions to the backscattered signal return. The null points on both sides of the main beams for the undesirable signal returns might not coincide with those for the signal return of interest. For example, from Table 4-2, the main beam for the HH polarizations signal return of interest, P_{rHH} , was defined by the sum of the prime polarization antenna gains, G_{tHM} and G_{rHM} . The main beams of the three undesirable signal returns (items 2, 3, and 4), on the other hand, were defined by a combination of prime and cross-polarization or by both cross-polarization antenna gains. By the observation of Figures 4-12 and 4-13, it became clear that the null points of the main beams of the signal return of interest and the undesirable signal contributions did not occur at the same cross-track angles. To make the estimate of the antenna coupling effect simple and reasonable, a single main beam boundary was applied to the calculations of all four components of desired and undesired signal returns. This boundary was defined by the null points of the main beam derived from the sum of G_{tHM} and G_{rHM} . The main beam

boundaries for the other three signal returns of different polarizations, P_{rHV} , P_{rVH} , and P_{rVV} (see Table 4-2), were defined in the same way.

The computer program was designed and implemented in accordance with Eq. (3-28). It evaluated the power outputs in terms of the square of the backscattered to calibrate signal voltage ratio, $(\frac{E_a}{E_c})^2$. Eight angles of incidence at 5° , 10° , 15° , 20° , 30° , 40° , 50° , and 60° were chosen for the computations. Since the along-track cell length is a fixed value of 50m, the bandwidth Δf_d changes with angle of incidence, decreasing from 88 Hz at $\theta = 5^\circ$ to 11 Hz at $\theta = 60^\circ$. Eight values of rolloff functions $Z_w(f_d)$ corresponding to the eight incident angles are entered in the program as a table. The values of the antenna feed line loss L and the calibrate constant K were given in Appendix A and were treated as constants in the program.

The program was coded in FORTRAN. It consisted of the main program and two subroutines. The first subroutine was used to obtain the values of the two-way antenna gain for the calculation of signal return, given the incident angle θ_o and cross-track angle ϕ_i . The second subroutine was used to evaluate the backscattering coefficient σ^0 , given the incident angle θ_i at area element ΔA_i . The functional dependence of σ^0_{VV} and σ^0_{HH} on θ_i is given in Eqs. (3-30) and (3-31) for water and land, respectively. The values of σ^0_{VH} and σ^0_{HV} were assumed to be 15 dB lower than σ^0_{HH} .

or σ_{VV}^0 at all angles of incidence. The program listing, flow chart, and the definition of variables are given in Appendix D. The program was run on the UNIVAC 1110 computer at JSC.

The two-way antenna gains for each of 16 combinations in Table 4-2, were previously saved on a magnetic tape, one file for each combination. Only the gain values within the main beam boundary corresponding to each of the four polarization groups were saved. For a σ^0 vs. θ curve, all the signal outputs corresponding to 16 components listed in Table 4-2 can be obtained with a single computer run.

5.2 THE σ^0 DEVIATION

As pointed out in Section 3, it is necessary to assume a functional dependence of σ^0 on the incident angle θ in the first method of σ^0 derivation. This assumption was made according to Eqs. (3-30) and (3-31) for the backscattering from water and land, respectively. The receiver signal outputs in terms of $(\frac{E_a^2}{E_c})$ from the computer runs are shown in Figures 5-1 and 5-2, respectively for the backscattering from water and land. These signal outputs could be compared with those actually obtained from observations to determine the validity of the assumed σ^0 dependence on θ . The procedure is straightforward and will not be performed here. Instead, a comparison will be made between the first (as represented by Eq. (3-28)) and second (as represented by Eq. (3-29)) methods of σ^0 derivation.

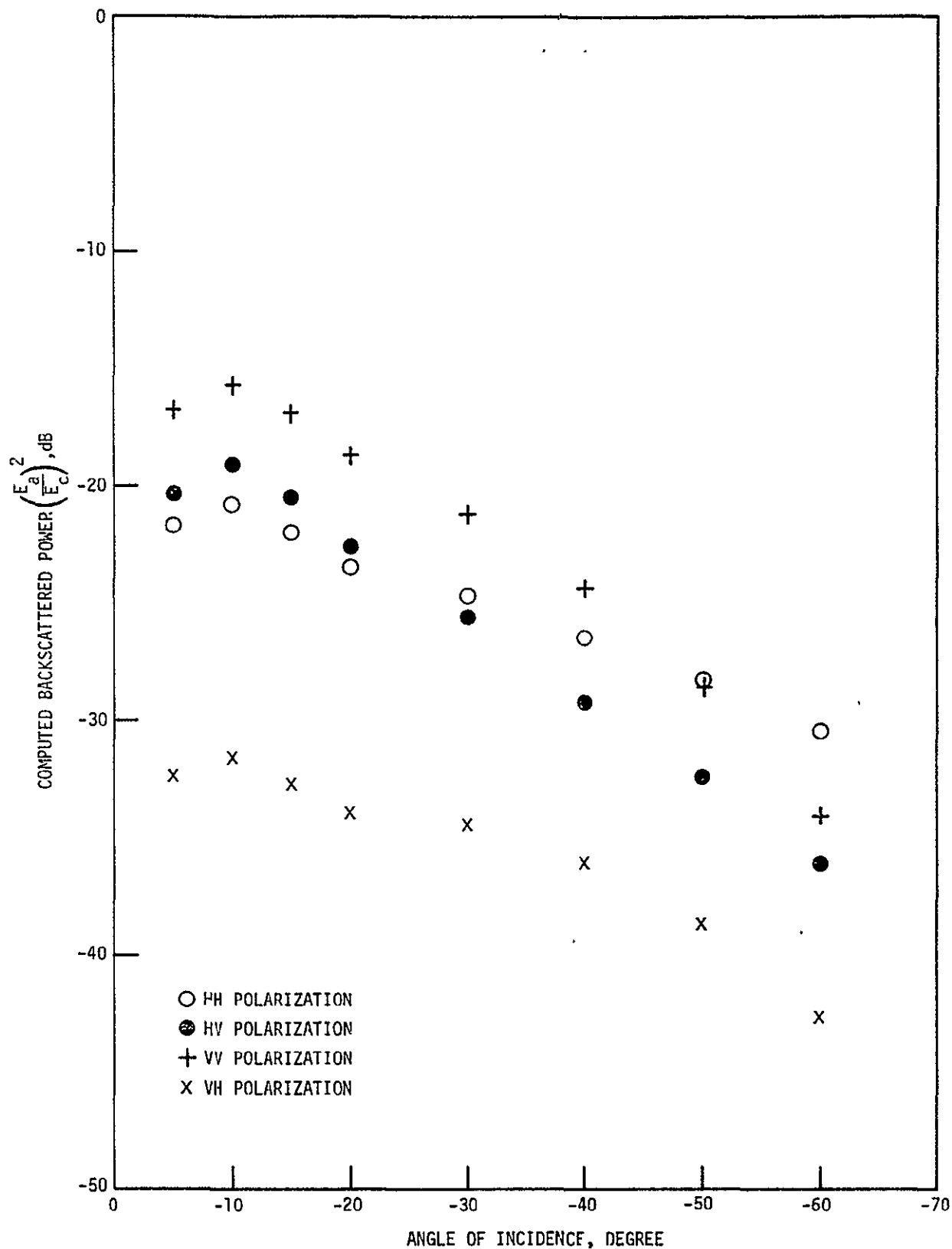


Figure 5-1.— The computed backscattered signals over water for all four polarization combinations.

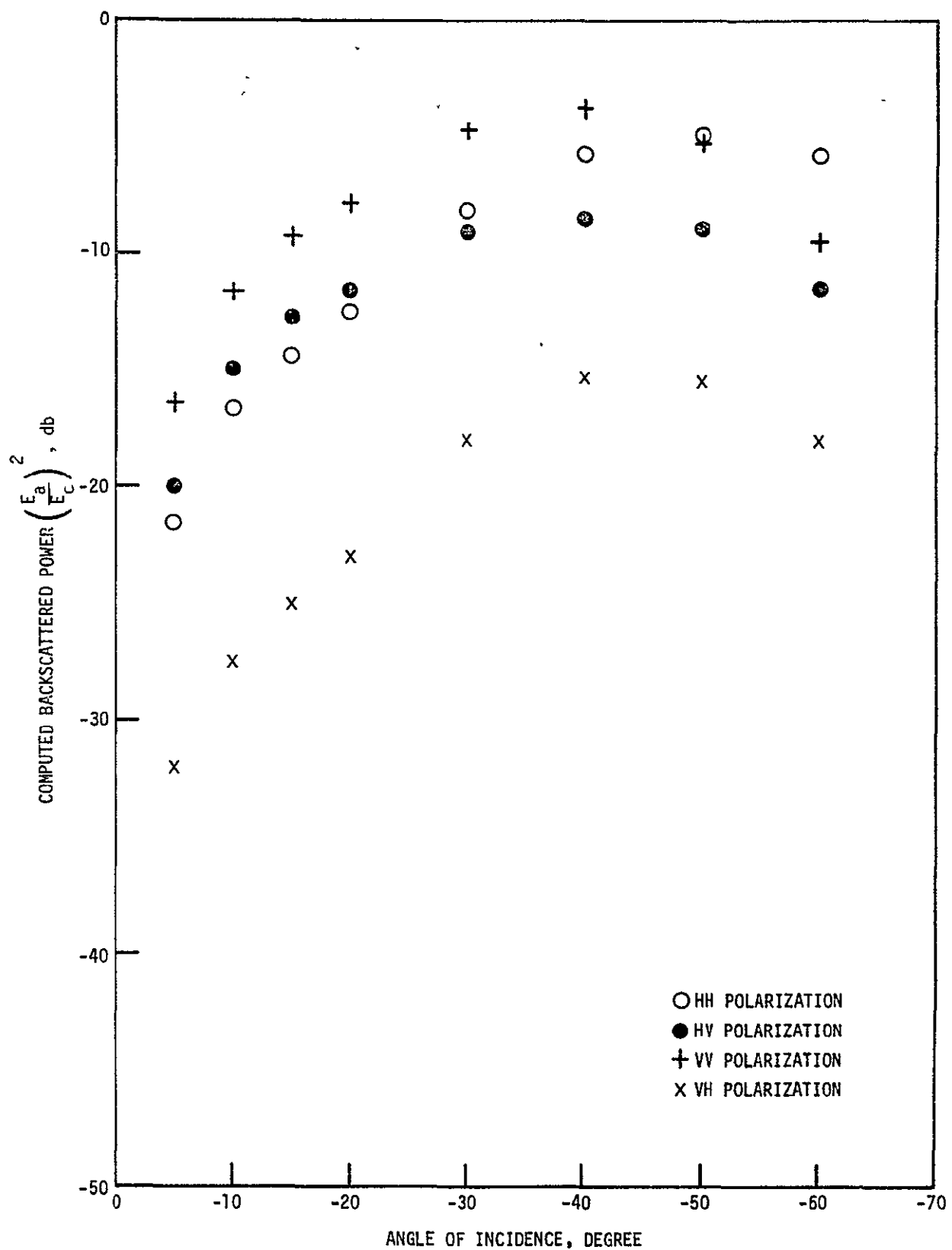


Figure 5-2.— The computed backscattered signals over land for all four polarization combinations.

To do such a comparison, the data displayed in Figures 5-1 and 5-2 were taken to be the receiver output signals for both methods. Then the σ^0 dependence on θ as derived from the first method will be given by Eq. (3-30) for water and Eq. (3-31) for land. The numerical values from these equations at the incident angles of -5° , -10° , -15° , -20° , -30° , -40° , -50° , and -60° degrees were listed under the " σ^0 " columns in Tables 5-1 and 5-2. The σ^0 's for both linear and cross polarizations over water were given in Table 5-1 while those over land in Table 5-2. Note that, as described in Subsection 5.1, no distinction is made between σ_{HH}^0 and σ_{VV}^0 and between σ_{VH}^0 and σ_{HV}^0 in this first method of σ^0 derivation.

The corresponding values of the σ^0 's derived from the second method of Eq. (3-29) were given in the same two tables under the columns of " σ_{HH}^0 ", " σ_{VV}^0 ", " σ_{HV}^0 ", and " σ_{VH}^0 ". The same numerical values for the parameters such as altitudes, aircraft speed, cell length, antenna feed line loss and calibrate constant K were used in Eqs. (3-28) and (3-29). It is clear from Table 5-1 that, for each of the four polarizations, the σ^0 derived from both methods are comparable for large $\theta \gtrsim |40^\circ|$. For small $\theta \lesssim |20^\circ|$, however, some differences are observed. At $\theta = -5^\circ$, for example, the σ^0 values derived from the second method could be $\sim 0.8 - 0.9$ dB smaller than those derived from the first method. The differences become smaller as the functional dependence of σ^0 on θ become less strong. This effect is clearly observed in Table 5-2 where the slope of

TABLE 5-1. - A COMPARISON OF σ^0 VALUES (dB) DERIVED FROM
TWO DIFFERENT APPROACHES - OVER WATER

Incident Angle θ (Degree)	Linear Polarization			Cross Polarization		
	σ^0	σ^0_{HH}	σ^0_{VV}	σ^0	σ^0_{HV}	σ^0_{VH}
- 5	+1.97	+1.17	+1.09	-13.03	-13.82	-13.80
-10	-2.64	-3.06	-3.01	-17.64	-18.49	-18.09
-15	-6.88	-7.08	-7.12	-21.88	-22.22	-22.22
-20	-10.76	-10.92	-11.03	-25.76	-25.94	-26.10
-30	-17.44	-17.32	-17.60	-32.44	-32.57	-32.63
-40	-22.68	-22.68	-22.77	-37.68	-37.68	-38.21
-50	-26.46	-26.43	-26.46	-41.46	-41.54	-41.56
-60	-28.80	-28.83	-28.88	-43.80	-43.81	-43.89

TABLE 5-2. - A COMPARISON OF σ^0 VALUES (dB) DERIVED FROM
TWO DIFFERENT APPROACHES - OVER LAND

Incident Angle θ (Degree)	Linear Polarization			Cross Polarization		
	σ^0	σ_{HH}^0	σ_{VV}^0	σ^0	σ_{HV}^0	σ_{VH}^0
-5	+1.58	+1.41	+1.36	-13.42	-13.52	-13.54
-10	+1.05	+0.99	+1.04	-13.95	-14.42	-14.02
-15	0.53	+0.57	+0.52	-14.47	-14.56	-14.55
-20	0	+0.02	-0.10	-15.00	-15.01	-15.14
-30	-1.05	-0.82	-1.12	-16.05	-16.08	-16.13
-40	-2.10	-2.04	-2.14	-17.10	-17.05	-17.57
-50	-3.15	-3.09	-3.12	-18.15	-18.19	-18.21
-60	-4.20	-4.22	-4.27	-19.20	-19.20	-19.28

σ^0 vs θ curve for land is much flatter than that for water. From this table, it is seen that, for any one of the four polarizations, the difference at $\theta = -5^\circ$ is only $\sim 0.1 - 0.2$ dB while the σ^0 values from the two methods for $\theta \gtrsim |40^\circ|$ remain comparable.

From these comparisons, it can be concluded that, when the slope of σ^0 vs θ curve is steep, the second method of σ^0 derivation tends to underestimate the σ^0 values. Although the beamwidth $\Delta\phi$ used in the second method was defined in terms of Eq. (4-1), it could be shown that the use of 3-dB beamwidth would not alter this conclusion. For example, it was pointed out in Section 4 that $\Delta\phi$ derived from 3-dB method is $\sim 1^\circ$ smaller than that derived from Eq. (4-1). If $\Delta\phi$ from Eq. (4-1) gives $\sim 8.5^\circ$ at $\theta = -5^\circ$ and 17° at $\theta = -60^\circ$, then the corresponding values from 3-dB approach would be $\sim 7.5^\circ$ and 16° , respectively. Those 3-dB $\Delta\phi$'s raise the σ^0 value by ~ 0.54 dB at $\theta = -5^\circ$ and ~ 0.26 dB at $\theta = -60^\circ$. Thus, using the 3-dB beamwidth in Eq. (3-29) improves σ^0 estimate at $\theta \lesssim |20^\circ|$, but overestimates the σ^0 value by a few tenths of 1 dB at $\theta \gtrsim |40^\circ|$.

5.3 THE EFFECT OF ANTENNA SIDE LOBES

The side-lobe contribution to the total power return can be estimated according to Eq. (4-4). To make a comparison of the power levels between the main-beam and the side-lobe signal returns, however, it is more convenient to write Eq. (4-4) in a form similar to Eq. (3-28). This is readily done and the final expression for the side-lobe signal return is:

$$\frac{Z_w(f_d)}{L K} \left(\frac{E_a}{E_c} \right)_S^2 \Delta f_d = \frac{\lambda^2}{(4\pi)^3} \sum_{i=1}^N \frac{G_i^2 \sigma_i^0}{R_i^4} \Delta x_i \Delta y_i \quad (5-1)$$

Here N is the number of the side-lobe area elements contributing to the power return at the Doppler frequency f_d and bandwidth Δf_d . $Z_w(f_d)$, L , and K have the same meaning and values as in Eq. (3-28). G_i^2 , σ_i^0 , Δx_i , and Δy_i were defined in Subsection 4.2. The variation of σ^0 with incident angle θ were again determined by Eqs. (3-30) and (3-31) for the backscattering from water and land, respectively. From Eq. (5-1), the side-lobe power level can be estimated in terms of $\left(\frac{E_a}{E_c} \right)^2$ and compared directly with main-beam power return shown in Figures 5-1 and 5-2.

The contributions of power return from the first side lobes on both sides of the main beams were computed and plotted in Figures 5-3 and 5-4 for water and land, respectively. The computations were made for all four polarizations. Clearly, the total side-lobe contributions are greater than those shown on the plots, because there are more than one side lobe on either side of the main beam for each of four polarizations. For instance, Figures 4-7 and 4-8 showed that there were about four or five side lobes on either side of any of four main beams. As a result, the side-lobe power contributions could be ~ 4 times (or ~ 6 dB) higher than those indicated in Figures 5-3 and 5-4 at $\theta = -20^\circ$.

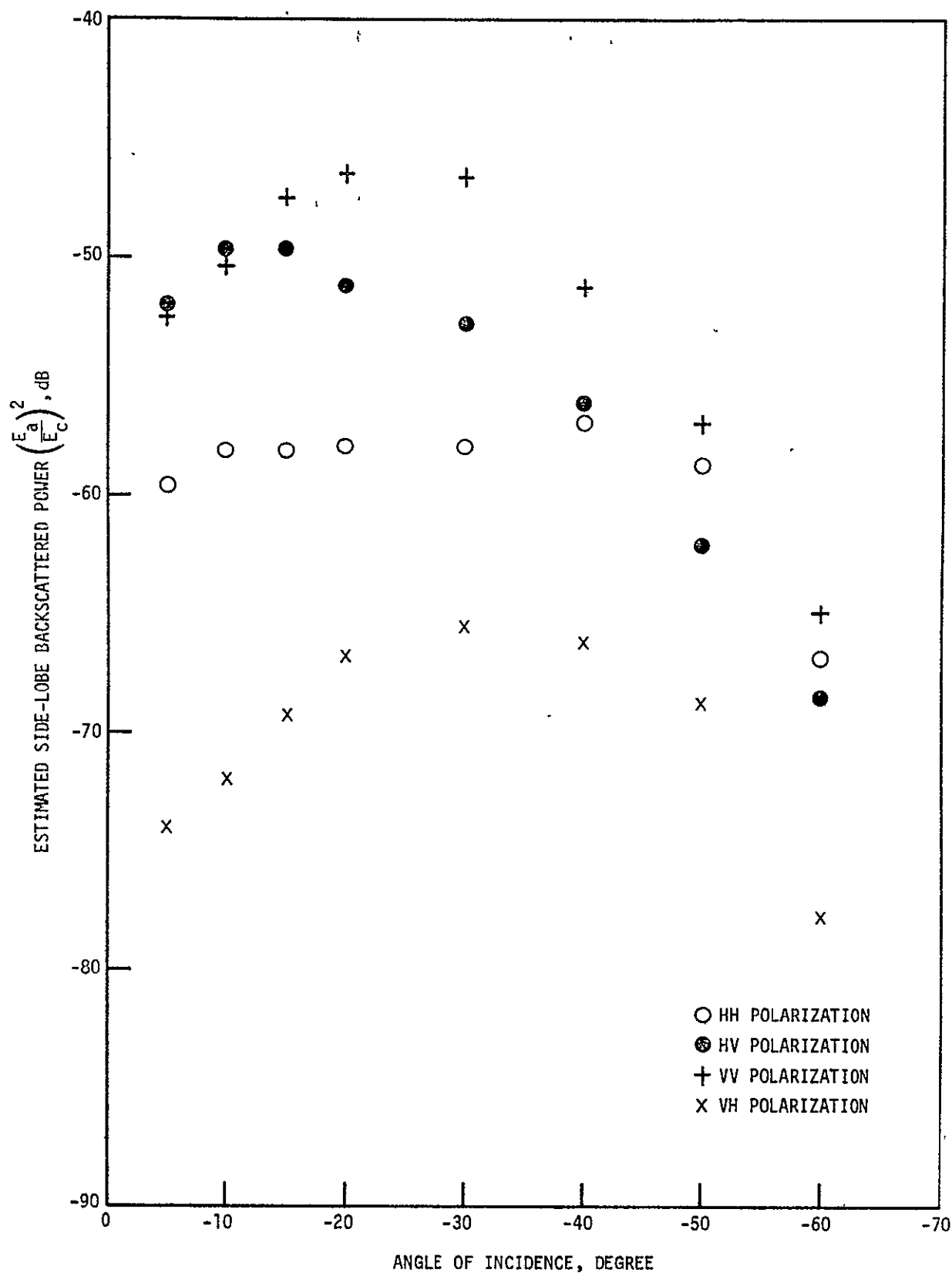


Figure 5-3.— Estimated first side-lobe signal return from water corresponding to the indicated main-beam incident angle.

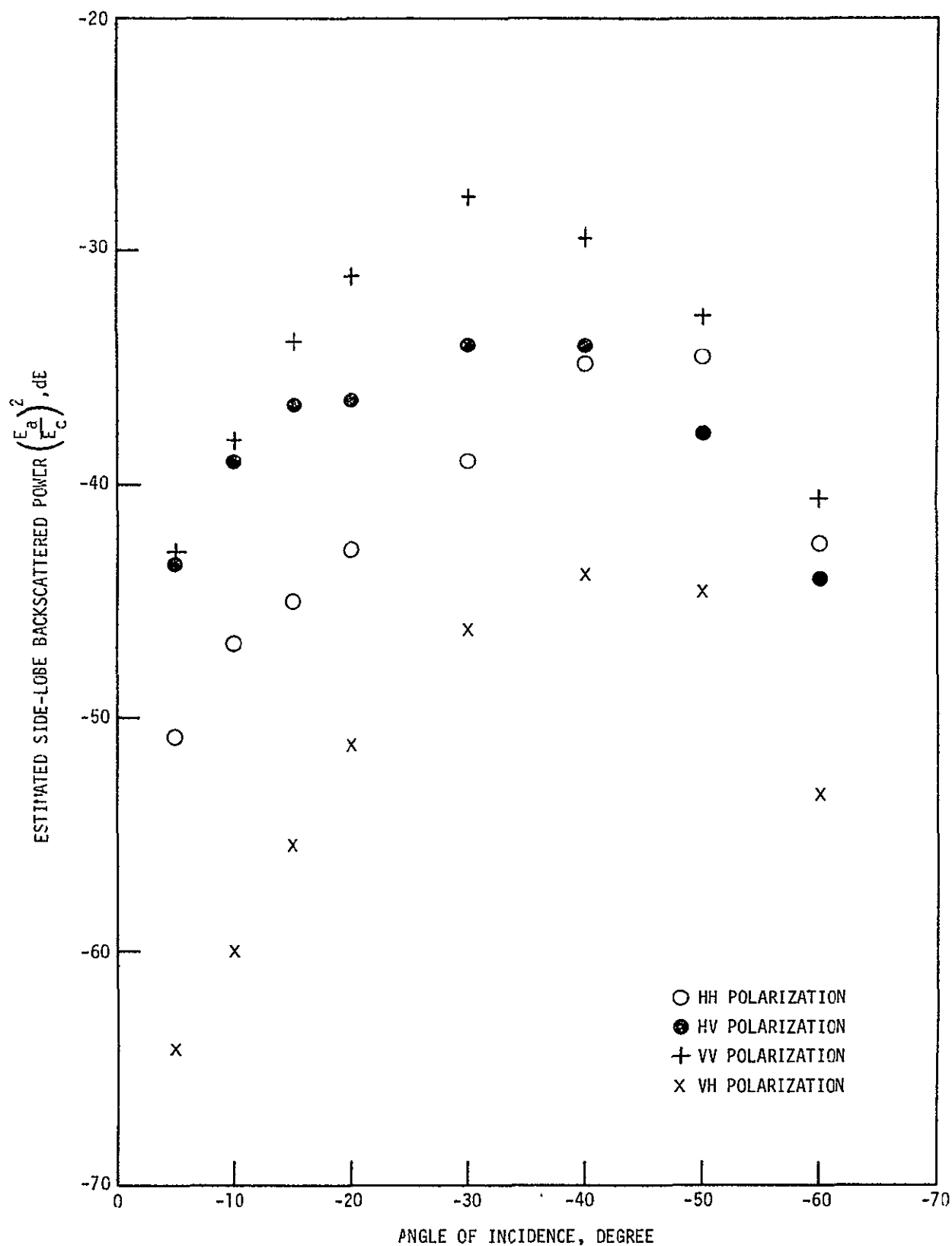


Figure 5-4.— Estimated first side-lobe signal return from land corresponding to the indicated main-beam incident angle.

Even if the side-lobe power levels shown in Figures 5-3 and 5-4 are raised by ~ 6 dB, their contribution to the total power is not significant enough to affect the σ^0 estimation. For example, with a 6-dB addition, the side lobe power level for VV polarization at $\theta = -20^\circ$ shown in Figure 5-3 would be ~ -40 dB. When compared with the corresponding backscattered power shown in Figure 5-1, the side-lobe power is found to be ~ 20 dB lower. This same effect is observed for the other three polarizations and for backscatterings from land (i.e., comparing Figures 5-2 and 5-4). Thus, it can be concluded that the side-lobe contributions to the backscattered signal powers are insignificant.

5.4 EFFECT DUE TO ANTENNA COUPLING

The computer outputs of the backscattered signals are displayed in sequential order in Tables 5-3, 5-4, 5-5, and 5-6 for HH, HV, VH, and VV polarizations. The computed signal outputs were derived for backscattering from both water and land and for incident angles at -5° , -10° , -15° , -20° , -30° , -40° , -50° , and -60° . The assignment of the backscattered signal components under columns "1", "2", "3", and "4" was the same as that given in Table 4-2. Consequently, the values under columns "1" were the backscattered powers of interest, while those under the other three columns were the backscattered powers due to antenna coupling. The values in the column "Total" gave the sums of those in the four columns. The values for σ_{HV}^0 and σ_{VH}^0 were taken to be 15 dB less than those for σ_{HH}^0 and σ_{VV}^0 in the computations.

TABLE 5-3. - NUMERICAL VALUES OF THE BACKSCATTERED SIGNALS
 (IN TERMS OF $\frac{E_a^2}{E_c}$) - HORIZONTAL TRANSMIT AND HORIZONTAL RECEIVE

Incident Angle θ (Degree)	HH (Water)					HH (Land)				
	1	2	3	4	TOTAL	1	2	3	4	TOTAL
- 5	-21.71	-59.86	-64.45	-70.68	-21.71	-21.47	-59.24	-63.61	-69.20	-21.47
-10	-20.69	-60.34	-62.98	-71.62	-20.69	-16.64	-56.17	-58.56	-66.92	-16.64
-15	-21.99	-61.83	-64.93	-73.99	-21.99	-14.34	-54.12	-56.92	-65.70	-14.34
-20	-23.40	-60.89	-67.05	-72.86	-23.40	-12.46	-49.89	-55.82	-61.16	-12.46
-30	-24.56	-60.63	-67.46	-71.19	-24.56	- 8.06	-44.07	-50.81	-54.01	- 8.06
-40	-26.49	-61.75	-68.23	-73.96	-26.49	- 5.85	-41.07	-47.51	-53.14	- 5.85
-50	-28.28	-62.77	-69.40	-73.37	-28.28	- 4.94	-39.40	-46.01	-49.92	- 4.94
-60	-30.39	-62.88	-70.31	-70.28	-30.38	- 5.78	-38.26	-45.68	-45.64	- 5.78

TABLE 5-4. - NUMERICAL VALUES OF THE BACKSCATTERED SIGNALS
 (IN TERMS OF $\frac{E_a^2}{E_c}$) - HORIZONTAL TRANSMIT AND VERTICAL RECEIVE

Incident Angle θ (Degree)	HV (Water)					HV (Land)				
	1	2	3	4	TOTAL	1	2	3	4	TOTAL
- 5	-20.35	-27.90	-26.85	-66.54	-18.89	-20.05	-27.12	-26.72	-66.22	-18.55
-10	-18.97	-28.13	-28.50	-68.86	-18.06	-14.90	-23.89	-24.43	-64.25	-13.97
-15	-20.45	-30.39	-32.25	-70.44	-19.78	-12.79	-22.69	-24.50	-61.67	-12.11
-20	-22.57	-30.31	-35.95	-71.02	-21.73	-11.64	-19.34	-24.90	-59.11	-10.79
-30	-25.58	-32.32	-41.03	-74.18	-24.64	- 9.09	-15.76	-24.37	-57.05	- 8.14
-40	-29.17	-35.37	-40.34	-71.78	-27.98	- 8.54	-14.71	-19.58	-50.97	- 7.33
-50	-32.40	-37.97	-37.26	-67.48	-30.35	- 9.05	-14.60	-13.87	-44.05	- 6.99
-60	-36.10	-39.74	-35.85	-63.60	-32.13	-11.49	-15.12	-11.23	-38.96	- 7.52

TABLE 5-5. - NUMERICAL VALUES OF THE BACKSCATTERED SIGNALS
 (IN TERMS OF $\frac{E_a^2}{E_c}$) - VERTICAL TRANSMIT AND HORIZONTAL RECEIVE

Incident Angle θ (Degree)	VH (Water)					VH (Land)				
	1	2	3	4	TOTAL	1	2	3	4	TOTAL
- 5	-32.38	-40.40	-45.43	-83.63	-31.56	-32.12	-39.99	-44.58	-82.11	-31.26
-10	-31.57	-38.33	-43.91	-80.71	-30.53	-27.50	-34.27	-39.47	-75.97	-26.45
-15	-32.72	-39.63	-45.44	-81.48	-31.73	-25.05	-31.92	-37.42	-73.25	-24.03
-20	-33.93	-41.65	-47.16	-82.94	-33.08	-22.97	-30.62	-35.91	-71.47	-22.10
-30	-34.52	-43.55	-47.51	-84.23	-33.82	-18.02	-26.99	-30.84	-67.15	-17.30
-40	-36.06	-45.22	-48.40	-86.29	-35.34	-15.42	-24.52	-27.67	-65.39	-14.69
-50	-38.81	-45.38	-50.34	-85.03	-37.70	-15.46	-21.99	-26.94	-61.57	-14.34
-60	-42.75	-45.17	-52.78	-81.72	-40.52	-18.14	-20.54	-28.16	-57.08	-15.90

TABLE 5-6. - NUMERICAL VALUES OF THE BACKSCATTERED SIGNALS
 (IN TERMS OF $\frac{E_a^2}{E_c^2}$) - VERTICAL TRANSMIT AND VERTICAL RECEIVE

Incident Angle θ (Degree)	VV (Water)					VV (Land)				
	1	2	3	4	TOTAL	1	2	3	4	TOTAL
- 5	-16.71	-55.13	-52.54	-59.91	-16.71	-16.44	-54.67	-52.30	-59.51	-16.44
-10	-15.69	-52.72	-54.31	-60.53	-15.69	-11.64	-48.68	-50.12	-56.15	-11.64
-15	-16.93	-54.18	-57.96	-64.02	-16.93	- 9.29	-46.51	-50.08	-55.76	- 9.29
-20	-18.76	-57.20	-61.44	-67.39	-18.76	- 7.83	-46.21	-50.25	-55.74	- 7.83
-30	-21.24	-61.08	-66.39	-73.22	-21.24	- 4.76	-44.55	-49.71	-56.25	- 4.76
-40	-24.39	-64.58	-64.75	-69.70	-24.39	- 3.76	-43.89	-44.00	-48.85	- 3.76
-50	-28.60	-66.30	-62.53	-65.32	-28.60	- 5.26	-42.91	-39.14	-41.88	- 5.26
-60	-34.05	-67.19	-62.99	-61.42	-34.03	- 9.44	-42.56	-38.37	-36.78	- 9.42

It is clear from Tables 5-3 and 5-6 that, by comparing columns "1" and "Total", the contribution to the backscattered signal due to antenna coupling is totally negligible in these cases. Therefore, it is not necessary to take the effect of antenna coupling into account when deriving σ_{HH}^0 or σ_{VV}^0 . The results shown in Tables 5-4 and 5-5, on the other hand, indicate an appreciable contribution to the total power. While there were non-negligible contributions at all eight incident angles under investigation, the most severe ones appeared to be at $\theta = -50^\circ$ and $\theta = -60^\circ$. For example, the values of the backscattered signals from water in Table 5-4 showed that the power contribution due to antenna coupling was $\sim 0.8 - 1.5$ dB and $\sim 2-4$ dB for $\theta \leq |40^\circ|$ and for $\theta \geq |50^\circ|$, respectively. This same remark could be made from the results of the backscattering from land in the same table. The values given in Table 5-5 showed the same pattern of variation with θ , although the effect was less severe.

The effect of the antenna coupling on the backscattered signal power level is expected to depend strongly on the relative values of σ_{HV}^0 and σ_{VH}^0 with respect to σ_{HH}^0 and σ_{VV}^0 . To make a rough estimate of this dependence, a similar computation was performed with σ_{HH}^0 and σ_{VV}^0 again given by Eqs. (3-30) and (3-31) for backscattering from water and land, respectively. But the values of σ_{HV}^0 and σ_{VH}^0 were assumed to be 10 dB less than those of σ_{HH}^0 and σ_{VV}^0 . The results of this computation are shown in Tables

5-7 and 5-8 respectively for HV and VH polarizations. For both water and land backscatterings the results in Table 5-7 indicated a ~ 1.7 dB contribution due to antenna coupling at $\theta = -60^\circ$. From $\theta = -5^\circ$ to $\theta = -15^\circ$, the contribution decreased from 0.5 dB to 0.2 dB. The values in Table 5-8 showed that the power contribution from antenna coupling was ~ 0.8 dB at $\theta = -60^\circ$ and fluctuated between 0.2 and 0.4 dB for other θ 's. The same computation was also made for HH and VV polarizations, and the results showed negligible contribution from antenna coupling, as to be expected from the observations of Tables 5-3 and 5.6. It can be shown that even if the values of σ_{HV}^0 and σ_{VH}^0 are comparable to those of σ_{HH}^0 and σ_{VV}^0 , the effect of antenna coupling remains negligible in σ_{HH}^0 and σ_{VV}^0 estimations.

From the above discussion, the following concluding remarks can be made. First, the derivation of σ_{HH}^0 and σ_{VV}^0 from data remotely acquired by the 1.6 GHz airborne scatterometer are free of antenna coupling effect. Secondly, if σ_{HV}^0 and σ_{VH}^0 are less than σ_{HH}^0 and σ_{VV}^0 by more than ~ 10 dB, the contribution of antenna coupling to the total backscattered powers in both HV and VH polarizations becomes appreciable. As a result of this undesirable contribution, σ_{HV}^0 and σ_{VH}^0 would be overestimated. Finally, the amount of the σ_{HV}^0 and σ_{VH}^0 overestimates depends on the relative values of σ_{HV}^0 and σ_{VH}^0 with respect to σ_{HH}^0 and σ_{VV}^0 . The results presented in Tables 5-4, 5-5, 5-7, and 5-8 indicate the highest overestimate to occur at incident angles ranging from -50° to -60° .

TABLE 5-7. - NUMERICAL VALUES OF THE BACKSCATTERED SIGNALS
(IN TERMS OF $\frac{E_a^2}{E_c}$) - HORIZONTAL TRANSMIT AND VERTICAL RECEIVE

Incident Angle θ (Degree)	HV (Water)					HV (Land)				
	1	2	3	4	TOTAL	1	2	3	4	TOTAL
- 5	-15.35	-27.90	-26.85	-61.54	-14.83	-15.05	-27.12	-26.72	-61.22	-14.52
-10	-13.97	-28.13	-28.50	-63.86	-13.66	- 9.90	-23.89	-24.43	-59.25	- 9.59
-15	-15.45	-30.39	-32.25	-65.44	-15.23	- 7.79	-22.69	-24.50	-56.67	- 7.56
-20	-17.57	-30.31	-35.95	-66.02	-17.29	- 6.64	-19.34	-24.90	-54.11	- 6.35
-30	-20.58	-32.32	-41.03	-69.18	-20.26	- 4.09	-15.76	-24.37	-52.05	- 3.77
-40	-24.17	-35.37	-40.34	-66.78	-23.76	- 3.54	-14.71	-19.58	-45.97	- 3.12
-50	-27.40	-37.97	-37.26	-62.48	-26.64	- 4.05	-14.60	-13.87	-39.05	- 3.28
-60	-31.10	-39.74	-35.85	-58.60	-29.42	- 6.49	-15.12	-11.23	-33.96	- 4.80

TABLE 5-8. - NUMERICAL VALUES OF THE BACKSCATTERED SIGNALS
 (IN TERMS OF $\frac{E_a^2}{E_c}$) - VERTICAL TRANSMIT AND HORIZONTAL RECEIVE

Incident Angle θ (Degree)	VH (Water)					VH (Land)				
	1	2	3	4	TOTAL	1	2	3	4	TOTAL
- 5	-27.38	-40.40	-45.43	-78.63	-27.10	-27.12	-39.99	-44.58	-77.11	-26.83
- 10	-26.57	-38.33	-43.91	-75.71	-26.22	-22.50	-34.27	-39.47	-70.97	-22.14
- 15	-27.72	-39.63	-45.44	-76.48	-27.38	-20.05	-31.92	-37.42	-68.25	-19.70
- 20	-28.93	-41.65	-47.16	-77.94	-28.64	-17.97	-30.62	-35.91	-66.47	-17.67
- 30	-29.52	-43.55	-47.51	-79.23	-29.29	-13.02	-26.99	-30.84	-62.15	-12.78
- 40	-31.06	-45.22	-48.40	-81.29	-30.82	-10.42	-24.52	-27.67	-60.39	-10.18
- 50	-33.81	-45.38	-50.34	-80.03	-33.43	-10.46	-21.99	-26.94	-56.57	-10.07
- 60	-37.75	-45.17	-52.78	-76.72	-36.91	-13.14	-20.54	-28.16	-52.08	-12.30

5.5 PRECISION IN THE OUTPUT SIGNAL

The discussion in this section is limited to those factors directly related to the 1.6 GHz scatterometer system only. The backscattered signal statistics which were discussed in great length elsewhere (ref. 11) will not be repeated here. The following gives a brief description for each of these factors considered to affect the performance of the scatterometer system.

5.5.1 ANTENNA PATTERN MEASUREMENTS

The uncertainty in the antenna pattern measurements was not recorded and supplied together with the gain values. M. Drexler (personal communication), of the New Mexico State University, placed this measurement uncertainty at $\sim \pm 0.5$ to ± 1 dB. An uncertainty of $\sim \pm 0.5$ to ± 1 dB in one-way gain results in the uncertainty of $\sim \pm 0.7$ to ± 1.3 dB for the two-way gain used in the σ^0 estimates. Another source of uncertainty comes from the fact that the gain measurements were made without a mockup to simulate the effect of the C130 aircraft. It is expected that, after the antennae were installed in the aircraft, the gain pattern would differ from that measured in the laboratory. The magnitude of this antenna pattern change cannot be estimated.

5.5.2 MEASUREMENTS OF CALIBRATE CONSTANT AND ROLLOFF FUNCTION

An uncertainty also exists in the measurements of the calibrate constant K and rolloff function $Z_{\omega}(f_d)$. A reasonable estimate placed this uncertainty to be no more than ± 0.5 dB for each of these measurements (S. C. Reid, personal communication).

5.5.3 AIRCRAFT PARAMETERS

Factors associated with aircraft motion such as roll, pitch, drift, vertical velocity, speed, and altitude changes could introduce errors in the received signals. The effects of roll, pitch, and vertical velocity are to alter the incident angle θ and were studied by Krishen et al. (ref. 4) with respect to the 13.3 GHz scatterometer system. Following the same approach, the errors introduced by these factors in the σ^0 estimates are presented in Tables 5-9, 5-10, and 5-11 for all four polarizations. In making these error calculations, the aircraft speed is assumed to be 77 m/sec, the vertical velocity to be 4 m/sec, and both roll and pitch angles to be 5° . Clearly, the change in θ due to pitch and vertical speed could be either plus or minus. Tables 5-10 and 5-11 show the case of larger modified incident angles only.

The changes of both aircraft speed and drift would introduce changes in the Doppler frequency associated with a given θ . For example, at $\theta = -30^\circ$, a change of speed from 77 m/sec to 70 m/sec would give a corresponding change in Doppler frequency from 411 Hz to 373 Hz. Similarly, a drift of $\sim 10^\circ$ would shift the Doppler frequency from 411 Hz to 405 Hz. These changes in Doppler frequency, if not corrected for in data processing, could result in erroneous σ^0 estimates.

The variation in aircraft altitude would also introduce error in the σ^0 estimate. From Eq. (3-3), σ^0 is observed to be proportional

TABLE 5-9. - ERRORS INTRODUCED IN THE σ^0 ESTIMATE
DUE TO AN AIRCRAFT ROLL OF 5°

Angle of Incidence (Degree)	Modified Angle of Incidence (Degree)	Error in σ^0 (dB)			
		HH	HV	VH	VV
- 5	- 7.07	0.35	-0.05	0.45	-0.02
-10	-11.17	0.30	0.35	0.17	-0.12
-15	-15.79	0.10	-0.05	-0.03	-0.25
-20	-20.59	0.09	-0.25	-0.31	-0.20
-30	-30.38	-0.04	-0.25	-0.05	-0.20
-40	-40.26	0.35	-0.24	0.09	-0.11
-50	-50.18	0.13	-0.12	0.20	-0.09
-60	-60.13	0	-0.15	0.12	-0.19

TABLE 5-10. - ERRORS INTRODUCED IN THE σ^0 ESTIMATE
DUE TO AN AIRCRAFT PITCH OF 5°

Angle of Incidence (Degree)	Modified Angle of Incidence (Degree)	Error in σ^0 (dB)			
		HH	HV	VH	VV
- 5	-10	0.70	0.20	0.85	0.83
-10	-15	0.40	0.55	0.14	0.25
-15	-20	-0.65	0.20	-0.85	-0.30
-20	-25	-1.44	-0.84	-1.60	-0.83
-30	-35	-1.53	-0.67	-1.32	-0.89
-40	-45	-1.18	-0.50	-0.51	0.23
-50	-55	-0.46	0.14	0.75	1.17
-60	-65	-0.48	0.16	0.39	0.95

TABLE 5-11. - ERRORS INTRODUCED IN THE σ^0 ESTIMATE
DUE TO AN AIRCRAFT VERTICAL SPEED OF 4 M/SEC

Angle of Incidence (Degree)	Modified Angle of Incidence (Degree)	Error in σ^0 (dB)			
		HH	HV	VH	VV
- 5	- 7.97	0.55	0	0.66	0.23
-10	-12.97	0.47	0.35	0.38	0.10
-15	-17.97	-0.15	0.20	-0.38	-0.30
-20	-22.97	-0.50	0.53	-0.74	-0.55
-30	-32.97	-0.68	-0.50	-0.64	-0.56
-40	-42.97	-0.76	-0.42	-0.33	-0.08
-50	-52.97	-0.28	0.07	0.48	0.60
-60	-62.97	0.05	-0.23	0.40	0.25

to R^4 and inversely proportional to the area A of the ground cell. Since R is proportional to altitude h and A to h^2 , σ^0 varies with h^2 . Thus, if h changes from 460 m to 440 m, the error in the σ^0 estimate would be ~ 0.39 dB.

5.5.4 ADDITIONAL ERROR SOURCES

Several additional error sources should also be considered. The first one is due to the antenna coupling discussed in Subsection 5.4. Care must be taken to eliminate or minimize this error source in estimating σ_{HV}^0 and σ_{VH}^0 from data remotely acquired by the 1.6 GHz airborne scatterometer system. The second source of error comes from the improper phase shift. This error source was discussed by Krishen et al. (ref. 4) in great detail. The results of their analysis were also applicable to the 1.6 GHz scatterometer system. The third source of error had its origin in the electronics assembly and was discussed in detail by Zuefeldt (ref. 9). This error source is rather small, however. Finally, if σ^0 vs. θ curve is steep, error could be present in σ^0 estimate at small incident angle, as pointed out in Subsection 5.1.

5.5.5 A BRIEF COMMENT

The uncertainty introduced in the antenna pattern measurements and by the presence of the aircraft body is difficult to correct. But this uncertainty can be treated as a bias and should not affect the repeatability of the data take from one mission to

another. The calibrate signal level could be adjusted to accommodate a particular mission. For each adjustment a new calibrate constant K would have to be measured and the measurement, therefore, should be made as precise as possible. The changes in aircraft parameters could introduce large error in σ^0 estimation. This type of error could be minimized by making corrections to the aircraft parameters during data processing. The error sources from improper phase shift, antenna coupling and steepness of σ^0 vs. θ curve could be reduced to a minimum by proper data reduction and manipulation. However, if the data quality is such that the ratio of the fore to aft signal powers is large or small compared to 1, a few degrees in phase shift error could result in large error in σ^0 estimate. The data should be used with reservation in this case.

6. CONCLUSIONS AND RECOMMENDATIONS

A system analysis was performed for the JSC 1.6 GHz airborne scatterometer. The principle of the scatterometer operation, calibration procedures, and data handling were described in some detail. The antenna gain patterns were studied in great length and the effects of the antenna cross polarization analyzed. Several error sources in the estimation of σ^0 were briefly discussed. In the course of the entire analysis, only the immediate power input to and the power output from the scatterometer system were dealt with. The interacting process of the RF waves with terrain surface was completely omitted. The terrain surface characteristics were reflected through the assumed functional forms of the back-scattering coefficient $\sigma^0(\theta)$.

The major conclusions that resulted from this study are listed in the following:

1. When σ^0 decreases rapidly with incident angle θ , which occurs in the backscattering from a quiet water surface, the normal approach of estimating σ^0 appears to give a low value at a small incident angle. If σ^0 vs θ curve over water is given by Figure 3-5, then the amount of the underestimate is ~ 0.8 dB at $\theta = 5^\circ$.
2. The gain patterns for the four antennae are quite different from one another. Most of the gain maxima occur at $\theta \approx \pm 50^\circ$. One of the two gain maxima for the vertical transmit antenna

occurs at $\theta \approx 15^\circ$ and the same for the vertical receive antenna at $\theta = -15^\circ$. The beamwidth variations with θ of all four antennae follow closely with one another.

3. The cross-polarization effects measured for the four antennae are found to have important bearing on the receiver signal outputs of both HV and VH polarizations. If σ_{HV}^0 and σ_{VH}^0 are less than σ_{HH}^0 and σ_{VV}^0 by ~ 10 dB, the power contributions due to antenna cross-polarization effects were found to raise the total power outputs at $\theta = 60^\circ$ by ~ 1.7 dB and ~ 0.8 dB for HV and VH receivers, respectively. The contributions become progressively larger as σ_{HV}^0 and σ_{VH}^0 become smaller compared to σ_{HH}^0 and σ_{VV}^0 .
4. The side-lobe contributions to the receiver signal outputs are negligible. For all cases examined, the backscattered power levels from the side lobes are ~ 15 dB or more below that from the main beam.
5. The variations of the aircraft parameters are found to affect the estimation of σ^0 . For example, a pitch of 5° could result in a ~ 1.5 dB error in some cases.

There are four recommendations that are considered to be important in the improvements of the system performance and the quality of the final results. They are listed below:

1. The second functional mode of operation (i.e., V transmit and H and V receive), which would give estimates of σ_{VV}^0 and σ_{VH}^0 , was seldom used in the past. The data outputs from a

few past missions indicated system instability in this mode of operation. If information provided by σ_{VH}^0 and σ_{VV}^0 are indispensable in the aircraft remote sensing of soil moisture, this problem will have to be fixed as soon as possible.

2. As indicated in the previous section, the effect of the aircraft body on the antenna pattern cannot be estimated. To improve the accuracy of σ^0 measurements, it is desirable to calibrate the aircraft scatterometer with a well calibrated ground system on a homogeneous terrain surface.
3. Data acquired from the 1.6 GHz scatterometer system should be checked with respect to the phase shift error. Corrections to the variations of the aircraft parameters should be made in the data processing.
4. The antenna cross-polarization effects should be taken into account in the estimations of σ_{HV}^0 and σ_{VH}^0 .

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APPENDIX A

THE CALIBRATION DATA

The laboratory calibration procedure for the 1.6 GHz scatterometer is almost the same as the 13.3 GHz scatterometer described by Reid (ref. 16). Therefore, only a simple description will be given wherever necessary. The calibrate data presented in the following were obtained during August of 1976 and were used in all the calculations in this document. Updating of these calibrate data are necessary for each JSME mission.

A.1 THE CALIBRATE CONSTANTS

As discussed in Section 3, the constant K which relates the transmit and receive signal levels to those actually recorded at the receiver output has to be predetermined in the laboratory. Figure A-1 shows a sketch of the setup for the measurement of this constant. The output power of the transmitter signal at 1.6 GHz was first measured by a power meter through a coupler. The signal was then attenuated by about 90 dB before entering a single side-band modulator. The frequency of the test signal to the receiver input became 1.6 GHz plus f_a , f_a being the frequency set by the audio-oscillator. f_a was chosen to be 5 KHz in the present case, where the land and sea filters had negligible effect of signal attenuation. The power level of this test signal was measured by a spectrum analyzer.

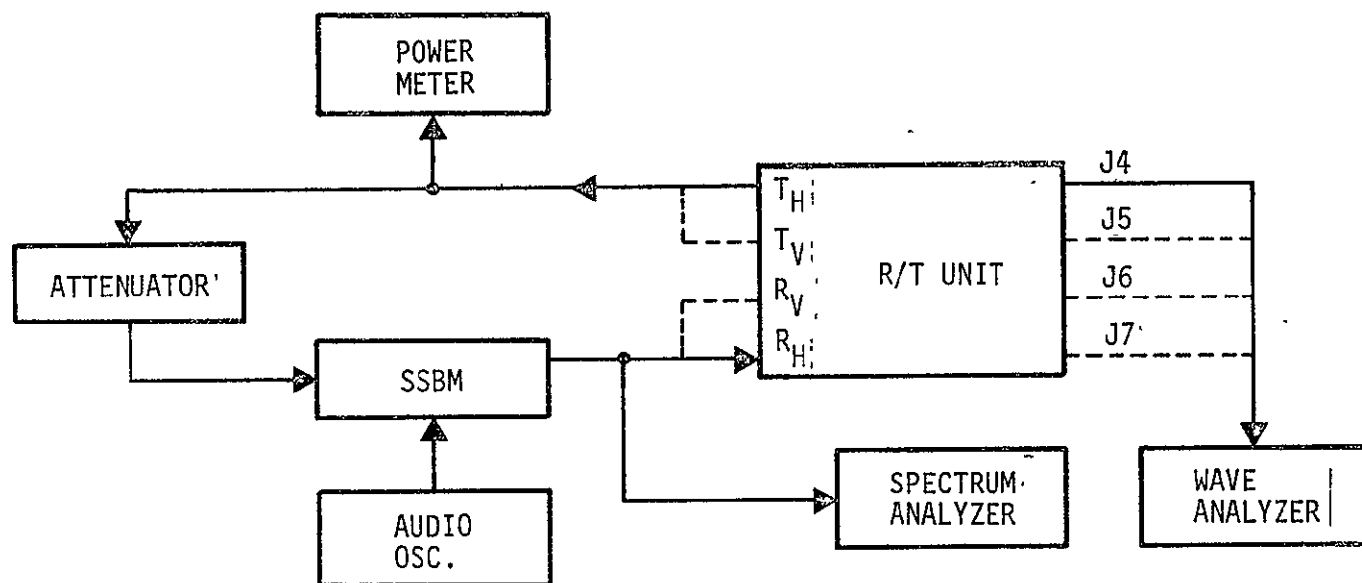


Figure A-1.— A sketch of the laboratory system calibration setup.

The routing of the calibrate signal at 1.9 KHz and the transmitter sample signal was not explicitly shown in Figure A-1, but was clearly displayed in Figure 3-3. The test signal at 5KHz and the calibrate signal at 1.9 KHz were separately measured by a wave analyzer at one of the four receiver output channels (J4, J5, J6 and J7). The calibrate signal was adjusted so that its output power level matched that of the test signal. Under this condition, the power difference between the transmit signal and the transmit sample signal and the amount of attenuation yielded the value for K. The calibrate signal level should not be changed until the next calibration measurement. There were four measured values for K depending on the polarizations of the transmit and receive RF waves. These were:

- 116.3 dB for horizontal transmit and horizontal receive
- 131.3 dB for horizontal transmit and vertical receive
- 118.8 dB for vertical transmit and vertical receive
- 119.3 dB for vertical transmit and horizontal receive.

A.2 VARIABLE GAIN RESPONSE TEST

The reason for this test was to verify that the changes in the output signal correspond to the changes in the variable gain amplifier settings. The setup and the procedure of this test were almost the same as those described in the subsection A.1 above. It was not necessary, however, that the receiver output signals at 1.0 KHz should be at the same level. The results of a typical test are summarized in Table A-1. The variable gain

TABLE A-1 The receiver output responses to the variable gain amplifier settings. All numbers are expressed in dB.

Variable Gain Setting Receiver Output Port (dB)		0	5	10	15	20
J4	CAL	-19.7	-14.8	- 9.8	- 4.8	+0.2
	TEST	-20.8	-16.0	-11.0	- 6.0	-1.0
J5	CODE	-20.8	-16.0	-11.0	- 6.0	-1.0
	TEST	-20.8	-16.0	-11.0	- 6.0	-1.0
J6	CAL	-34.0	-29.0	-24.0	-19.0	-14.0
	TEST	-20.8	-15.8	-10.8	- 5.8	-0.8
J7	CAL	-33.3	-28.5	-23.5	-18.5	-13.5
	TEST	-20.8	-15.8	-10.8	- 5.8	-0.8

amplifier setting was successively changed from 0 to 20 dB in 5-dB steps. The signals at all four receiver output ports J4, J5, J6 and J7 were observed to respond in a 5-dB increment as expected. The terms CAL, TEST, and CODE refer to the calibrate, test, and coding (at 2.1 KHz) signals respectively.

A.3 THE ROLLOFF FUNCTION

The measurements of the sea and land filter responses as a function of audio-frequency (rolloff functions) were again similar to the procedure described in Subsection A.1. After the values of K were determined, for example, the audio-frequency of the test signal could be reduced and the output test signal power level measured in reference to the calibrate signal power level. The results of these measurements are summarized in Table A-2 for both land and sea filters. It is clear from this table that, for either land or sea filter, the frequency responses of all four receiver output channels are very similar. This indicates a good gain balance among all four receiver channels. The average frequency responses of all four receiver channels for both land and sea filters were plotted in Figure 3-4 and used in the calculations of the receiver output signal levels.

A.4 ANTENNA CABLE LOSSES

The measured K values did not include the signal attenuation losses due to the co-axial cables from the R/T unit to the antennae and back to the R/T unit. These losses were measured independently for each of the four combinations of polarizations.

TABLE A-2
THE MEASURED RESPONSES (ROLLOFF FUNCTIONS) IN dB FOR THE
1.6 GHZ SCATTEROMETER RECEIVER OUTPUTS

TEST FREQ. (Hz)	LAND FILTER ON				SEA FILTER ON			
	J-4	J-5	J-6	J-7	J-4	J-5	J-6	J-7
1800	+0.5	+0.25	+0.25	+0.25	0	0	0	0
1600	+0.25	+0.25	+0.25	+0.25	0	0	0	0
1400	+0.3	+0.1	+0.2	+0.2	0	+0.2	-0.1	-0.1
1200	+0.3	+0.1	+0.2	+0.2	-0.2	-0.25	-0.2	-0.2
1000	+0.4	+0.1	+0.2	+0.2	-0.6	-0.8	-0.7	-0.7
800	+0.5	+0.1	+0.3	+0.2	-1.0	-1.3	-1.1	-1.1
700	+0.5	+0.1	+0.2	+0.2	-1.5	-1.8	-1.6	-1.6
600	+0.5	+0.1	+0.3	+0.2	-2.0	-2.2	-2.1	-2.1
500	+0.5	+0.1	+0.3	+0.2	-3.0	-3.4	-3.1	-3.1
400	+0.3	+0.1	+0.2	+0.2	-4.0	-4.2	-4.0	-4.0
300	+0.4	+0.1	+0.2	+0.2	-5.7	-6.0	-5.7	-5.8
200	+0.1	0	0	0	-9.0	-9.0	-9.0	-9.0
100	-0.5	-0.9	-0.5	-0.7	-15.0	-15.5	-15.0	-15.0
85	-1.0	-1.0	-1.0	-1.0	-17.0	-17.0	-17.0	-17.0
70	-1.5	-1.8	-1.5	-1.5	-19.0	-19.0	-19.0	-19.0
60	-2.0	-2.5	-2.5	-2.0	-20.0	-20.0	-20.5	-21.0
50	-3.0	-3.0	-3.0	-3.5	-23.5	-23.5	-23.0	-23.5
40	-4.5	-4.5	-4.0	-4.5	-27.0	-26.5	-26.0	-27.0
35	-5.5	-5.5	-5.5	-6.0	-29.0	-29.0	-28.5	-28.5

The results were:

1.9 dB for horizontal transmit and horizontal receive,
1.8 dB for horizontal transmit and vertical receive,
1.9 dB for vertical transmit and vertical receive,
2.0 dB for vertical transmit and horizontal receive.

APPENDIX B

THE PIN DIODE MODULATOR

To derive the absolute value of the backscattering coefficient σ^0 as a function of incident angle θ^i from the backscattered Doppler signal return, it is necessary to maintain a reference calibrate signal, proportional to the transmitted power, in the receiver output. The 1.6 GHz scatterometer system utilizes the Hewlett Packard 5082-3040 PIN diode modulators to provide the reference signal required in the data reduction. The method is to amplitude modulate a sample of the transmit power by a calibrate signal of 1.9 KHz and pass the sidebands of the modulation through the receiver network along with the backscattered signal. As shown in Figure 2-2, two PIN diode modulators, one for the horizontal receiver (associated with H antenna) and the other for the vertical receiver (associated with V antenna), are used in the system. Only the two data channels which are 90° phase-shifted contain the calibrate signals.

A PIN diode is a silicon semiconductor consisting of a layer of high resistivity intrinsic material (I) sandwiched between a p-type layer on one side and a n-type layer on the other. At low frequencies $< \frac{1}{\tau}$, where τ is the charge recombination lifetime in the I layer, it behaves as a normal rectifier. At frequencies $> \frac{1}{\tau}$, it appears as a linear resistance, whose value depends mainly on the conductivity and the geometry of the I

layer (refs. 17, 18 and 19). Approximately, the diode resistance at these high frequencies is proportional to the square of the I layer thickness and inversely proportional to the product of the bias current I_b and τ . As the diode is shunt connected across the RF signal path and forward biased, it acts as an attenuator with attenuation in the circuit given by (ref. 19)

$$\alpha_a = 20 \log \left(1 + \frac{Z_0}{2R_I} \right) \quad (B-1)$$

where Z_0 is the load resistance and R_I is the diode RF resistance at a specified bias current. Since R_I is inversely proportional to I_b approximately, α_a increases with I_b . A typical attenuation of a HP 5082-3040 PIN diode at RF frequency of ~8 GHz is shown in Figure B-1 as a function of I_b (ref. 20). For this type of PIN diodes, τ is typically ~ 200 nsec. At the frequency of 1.6 GHz $\gg \frac{1}{2\pi\tau}$, the same general dependence of α_a is expected. It is clear from this curve that the direct maximum attenuation of RF power obtainable from a single PIN diode of this kind is ~25-30 dB.

If a small-amplitude AC current is added to the direct DC bias current, then it is possible to obtain a calibrate signal at about the same power level to the backscattered signal. In this case, the bias current can be written as

$$I_b \sim \alpha(1 + \gamma \cos \omega_c t) \quad (B-2)$$

Here $\omega_c = 2\pi f_c$ with $f_c = 1.9$ KHz and $0 < \gamma < 1$. γ can be adjusted so that best sensitivity of the scatterometer system is obtained

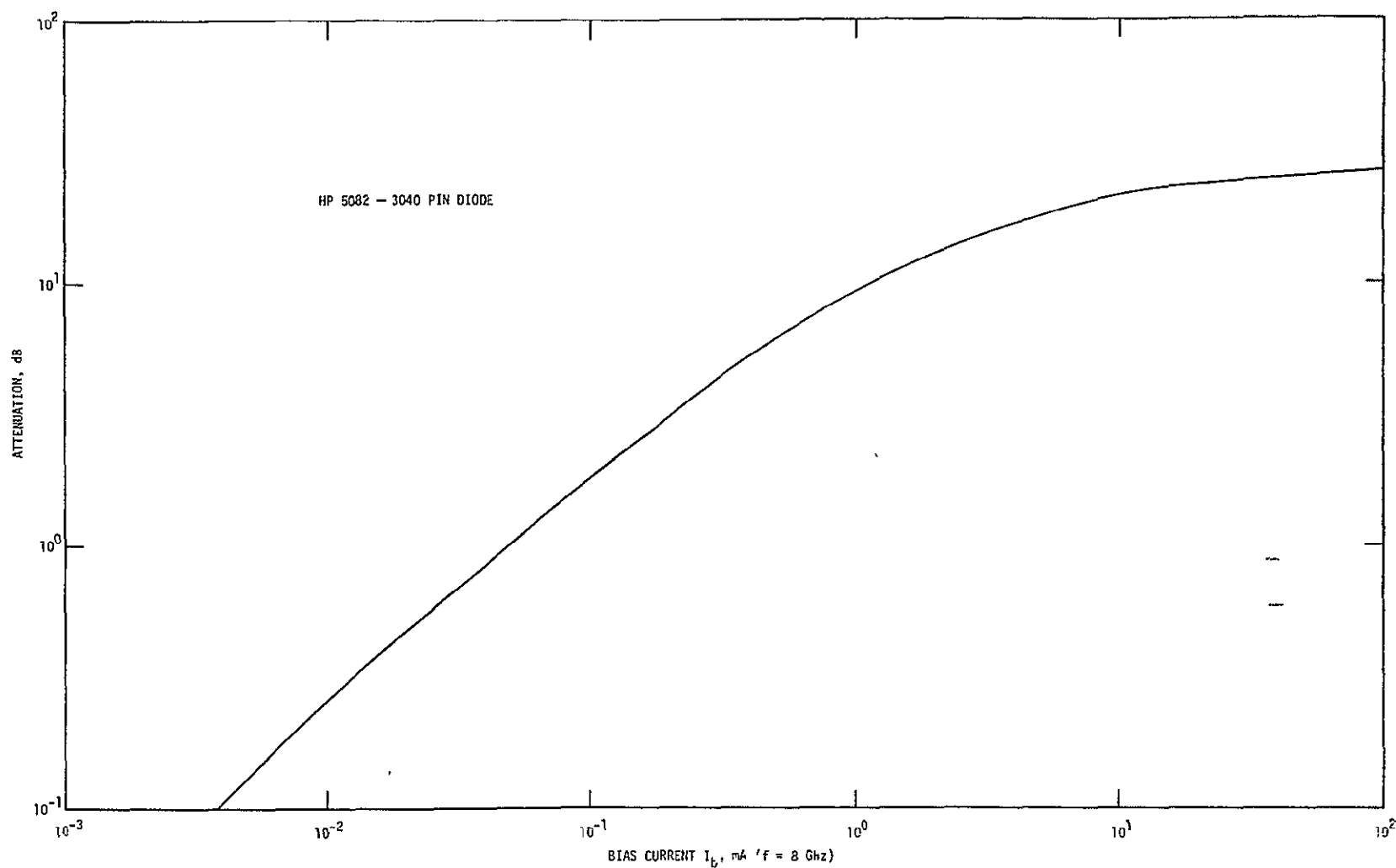


Figure B-1.— A typical attenuation above zero bias insertion loss versus bias current for a 5082 - 3040 PIN diode (ref. 20).'

without signal saturation. When the transmit sample signal is modulated by the signal source expressed in Eq. (B-2), the resultant output is:

$$\alpha P_2 V_0 \cos \omega_0 t (1 + \gamma \cos \omega_0 t) \quad (B-3)$$

where $P_2 V_0 \cos \omega_0 t$ represents the transmit sample signal voltage. The first term in Eq. (B-3) is filtered out after the mixer and the subsequent amplifier stages. The second term provides the desired calibrate signal in the data reduction.

The characteristics of the calibrate signal modulators for the 1.6 GHz scatterometer is given in Table B-1, while that for the HP 5082-3040 PIN diode is given in Table B-2.

TABLE B-1
CHARACTERISTICS OF CALIBRATE SIGNAL MODULATORS
FOR 1.6 GHz SCATTEROMETER

TYPE OF CIRCUIT	PIN diode modulator
RF carrier	1.6 GHz low level signal from transmitter
Modulator waveform	1.9 ± 0.06 KHz sine wave
Waveform synchronization	Free running
Type of modulation	A-M
Modulation depth	Adjustable
Number required and location	Two total, one for each 90° phase shifted channel.
Level of equivalent 2 KHz Doppler signal at aircraft speed of ~160 knots	~-140 dB to -160 dB relative to transmitter power.

TABLE B-2
PIN DIODE CHARACTERISTICS

Type	HP 5082 - 3040
Carrier lifetime	~200 nsec
VSWR at 1.6 GHz	~1.1
Insertion loss at 1.6 GHz	~0.2 dB
Maximum attenuation at 1.6 GHz	30 dB
Operating temperature range	-65°C to +125°C

APPENDIX C THE MIXER MATHEMATICAL MODEL

A mixer is essentially a nonlinear resistive element which combines the input signal with the local oscillator signal (the transmitter sample signal in the present case) and outputs signals at the frequencies harmonically related to the input and local oscillator frequencies and their mixtures. The characteristics of the nonlinear resistance for a typical mixer is shown in Figure C-1. The transmitter sample signal is given by Eq. (3-15):

$$S_T(t) = \rho_2 V_O \cos \omega_O t \quad (C-1)$$

and the combination of the backscattered and the calibrate signals in channel 2 can be written as:

$$S_C(t) = \frac{k_2}{\sqrt{2}} \left\{ -C_f \sin \left[(\omega_O + \omega_d)t + \phi_f \right] - C_a \sin \left[(\omega_O - \omega_d)t + \phi_a \right] \right\} \\ + \alpha \rho_2 V_O \cos \omega_O t (1 + \gamma \cos \omega_O t) \quad (C-2)$$

The power series representation of the current I_{2m} flowing in the mixer may be expressed in terms of the voltage across the mixer terminals (ref. 21) as:

$$I_{2m} = a_0 + a_1 V + a_2 V^2 + a_3 V^3 + \dots + a_n V^n \quad (C-3)$$

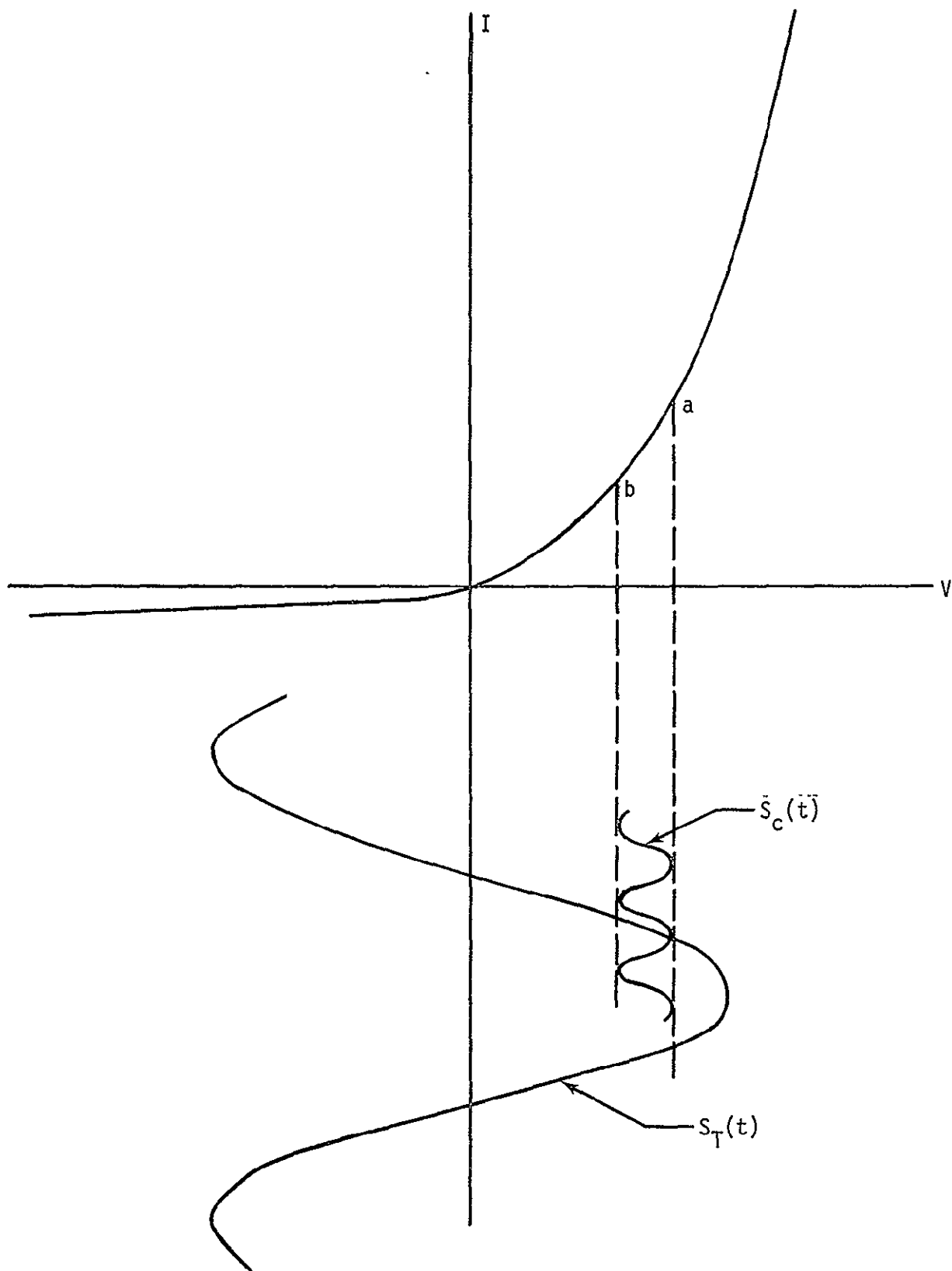


Figure C-1.— Characteristics of a nonlinear resistive element.

with

$$V(t) = S_T + S_C$$

The coefficient a_n generally decreases rapidly with n . Neglecting terms with $n \geq 3$, Eq. (A-3) becomes

$$I_{2m} \approx a_0 + a_1 (S_T + S_C) + a_2 (S_T + S_C)^2 \quad (C-4)$$

The signal given by Eq. (C-4) consists of DC terms as well as terms at the frequencies of ω_d , ω_c , $\omega_o \pm \omega_d$, $2\omega_o$, $\omega_o \pm \omega_c$, $2\omega_o \pm \omega_d$, $2\omega_o \pm 2\omega_d$, $2\omega_o \pm 2\omega_c$, and $2\omega_o \pm \omega_c$. For the 1.6 GHz scatterometer system, $\omega_o = 1.6$ GHz, $\omega_c = 1.9$ KHz and $\omega_d \leq 1$ KHz for aircraft speed of ≤ 200 knots. All except the terms at audio-frequencies of ω_d and ω_c are filtered out in the subsequent amplifier stages. Substituting the expressions for S_T and S_C as given by Eqs. (C-1) and (C-2) into Eq. (C-4) and retaining only the terms at the frequencies $\leq \omega_c$, the output signal voltage across the mixer load register R_2 is

$$\begin{aligned} V_{2m} = a_2 R_2 & \left\{ - \frac{k_2}{\sqrt{2}} C_f \propto \rho_2 V_o \sin (\omega_d t + \phi_f) \right. \\ & + \frac{k_2}{\sqrt{2}} C_a \propto \rho_2 V_o \sin (\omega_d t - \phi_a) \\ & \left. - \gamma \propto^2 \rho_2^2 V_o^2 \cos \omega_c t + C_f C_a \sin (2\omega_d t + \phi_f - \phi_a) \right\} \end{aligned}$$

$$\begin{aligned}
& + \frac{\alpha \gamma C_f \rho_2 V_o}{2} \sin \left[(\omega_c - \omega_d)t - \phi_f \right] \\
& - \frac{\alpha \gamma C_a \rho_2 V_o}{2} \sin \left[(\omega_c - \omega_d)t - \phi_a \right] \} \quad (C-5)
\end{aligned}$$

For an altitude of $\sim 457\text{m}$, an aircraft speed of $\sim 77\text{m/sec}$, σ^0 of $\sim -30\text{ dB}$, and the two-way antenna gain of $\sim 20\text{ dB}$, the power level of the input signal level at the receiver was estimated to be $\sim 116\text{ dBm}$. α can be adjusted such that $k_2 \alpha \rho_2 V_o \gg C_f, C_a$. γ is adjustable and is generally much less than 1 (see Appendix B). Consequently, the last three terms in Eq. (C-5) can be neglected compared to the first three terms. Let $\beta_2 = a_2 R_2 \alpha$, then Eq. (C-5) becomes

$$\begin{aligned}
V_{2m} = \frac{k_2}{\sqrt{2}} \beta_2 \rho_2 V_o \left[-C_f \sin (\omega_d t + \phi_f) + C_a \sin (\omega_d t - \phi_a) \right. \\
\left. - \gamma \alpha \rho_2^2 V_o^2 \cos \omega_c t \right] \quad (C-6)
\end{aligned}$$

The signal voltage at the mixer output in channel 1 can be derived by the same procedure. The result is given by:

$$V_{1m} = \frac{k_1}{\sqrt{2}} \beta_1 \rho_1 V_o \left[C_f \cos (\omega_d t + \phi_f) + C_a \cos (\omega_d t - \phi_a) \right] \quad (C-7)$$

APPENDIX D

THE COMPUTER PROGRAM FLOW CHART AND LISTING

The listing and the flow charts of the computer program used in the numerical calculation are given in this appendix. The flow charts of the main program and two subroutines, BCOFF and GAT, are sequentially displayed in Figures D-1, D-2, and D-3. The listing of the whole program is given in Table D-1.

Three subroutines RINIT, RWRITE, and RREAD are used to initialize, to write data into, and to read data out of the high speed drum storage. These subroutines are stored in the local computer programs library and explained in the JSC CAD Procedure Manual, Exec 8 System General User Information.

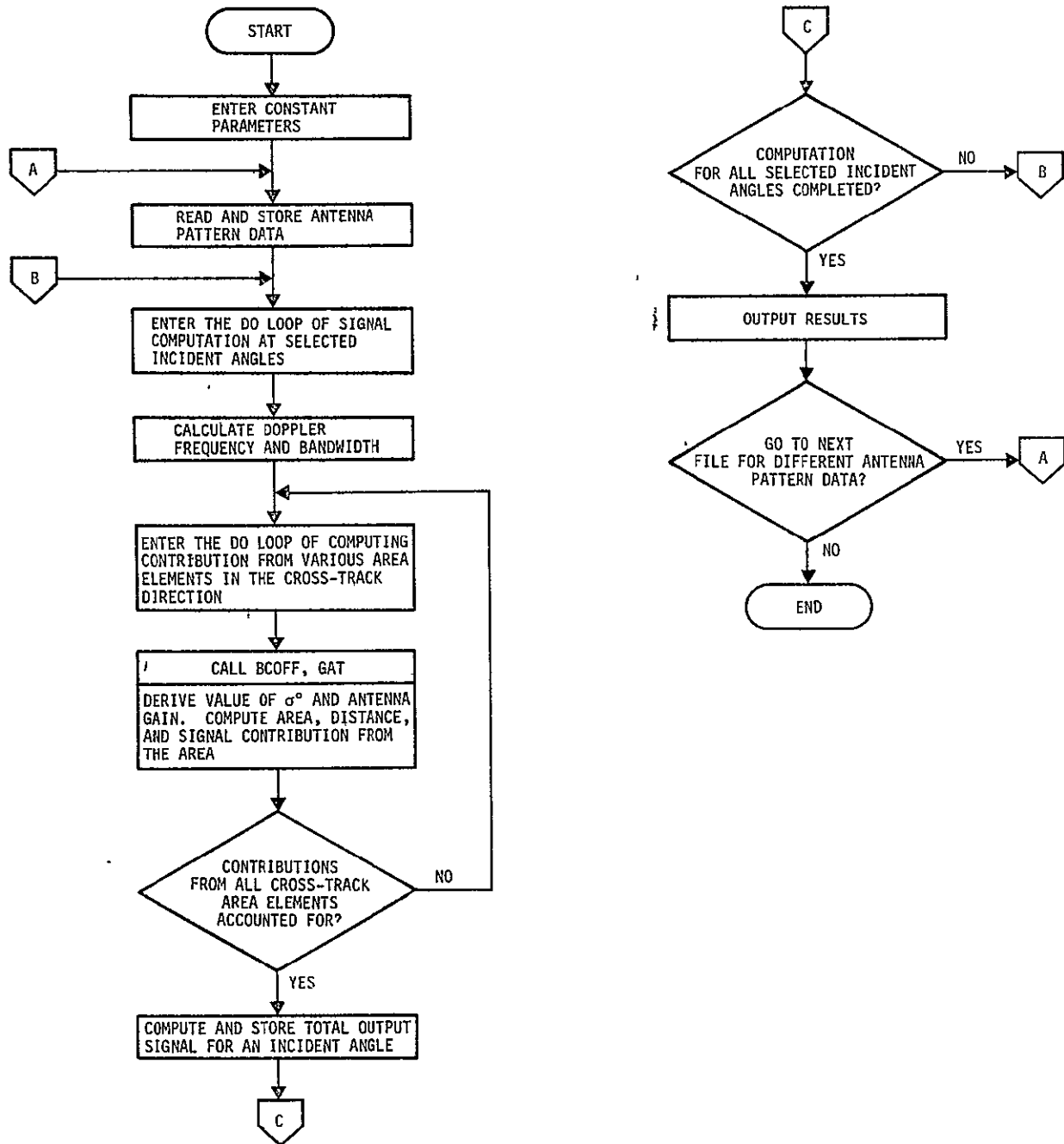


Figure D-1.— The MAIN program flow chart.

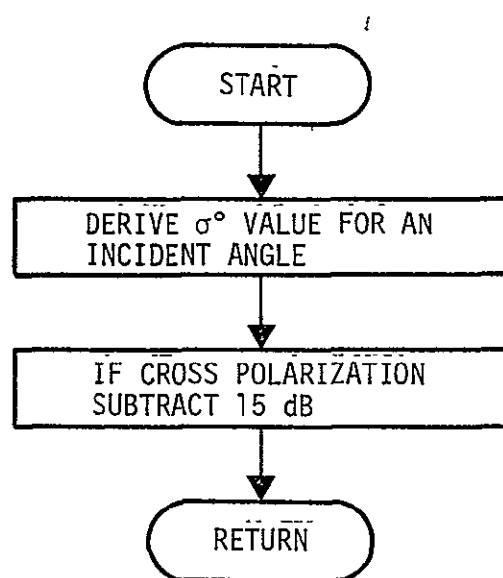


Figure D-2.— Subroutine BCOFF flow chart.

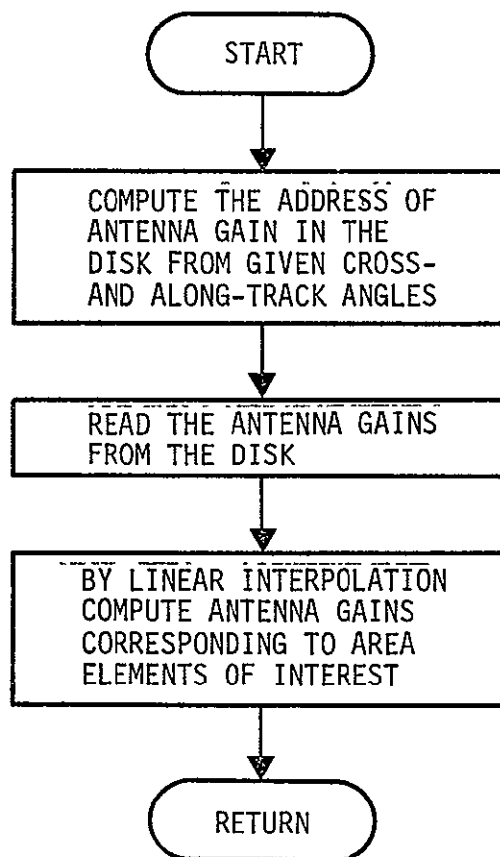


Figure D-3.- Subroutine GAT flow chart.

TABLE D-1. - THE COMPUTER PROGRAM LISTING

```

1*      DIMENSION THETA(8),POW(8),ROFF(8),BUF(92)
2*      DIMENSION ALOSS(4),SCAL(4)
3*      DATA THETA/5.,10.,15.,20.,30.,40.,50.,60./
4*      DATA ROFF/17.8,11.8,8.6,6.7,4.0,2.7,1.8,1.4/
5*      DATA ALOSS/1.9,1.8,2.0,1.9/
6*      DATA SCAL/116.3,131.3,119.3,118.8/
7*      COMMON/B/IDADD,IX,IY,IXMAX,IYMAX
8*      C ENTER CONSTANT PARAMETERS.
9*      PIE=3.1416
10*      HA=460.
11*      VA=77.
12*      SLAMD=0.1875
13*      CELEN=50.
14*      PI=PIE/180.
15*      IFIL=1
16*      IX=92
17*      IY=36
18*      IXMAX=92
19*      IYMAX=36
20*      CALL RINIT(IDADD,NWRDS)
21*      DO 600 M=1,4
22*      NS=0
23*      IEOF=IFIL+4
24*      50 CONTINUE
25*      NS=NS+1
26*      C GET DISC AREA
27*      K=IDADD
28*      C PUT ANTENNA FOOTPRINT ON DISC
29*      100 READ(1,4000,END=200)(BUF(NR),NR=1,92)
30*      C WRITE(6,5000)(BUF(NR),NR=1,92)
31*      CALL RWRITE(K,BUF,IX,NSTAT)
32*      5 IF(NSTAT.EQ.1) GO TO 5
33*      K=K+IX
34*      GO TO 100
35*      200 CONTINUE
36*      C ENTER MAIN LOOP OF COMPUTATION.
37*      C ALONG-TRACK COMPUTATION.
38*      DO 500 I=1,8
39*      THTA=THETA(I)*PI
40*      FDOP=2.*VA*SIN(THTA)/SLAMD
41*      DF=2.*VA*(COS(THTA)**3)*CELEN/HA/SLAMD
42*      CLAMD=SLAMD**2
43*      CPI=(4.*PIF)**3
44*      SUM=0.
45*      Q=TAN(THTA)**2
46*      C CROSS-TRACK COMPUTATION.
47*      DO 400 J=1,21
48*      PHI=(J-1)*PI*2.
49*      PHIL=((J-1)*2.-1.)*PI
50*      PHIR=((J-1)*2.+1.)*PI
51*      P=TAN(PHI)**2.
52*      DX=HA*(TAN(PHIR)-TAN(PHIL))
53*      DY=CELEN*SQRT(1+P)
54*      DA=DX*DY
55*      R=HA*SQRT((1+P)*(1+Q))
56*      RR=R**4.
57*      THINC=ACOS(HA/R)
58*      THINC=THINC/PI
59*      CALL BCOFF(THINC,SIGMA,NS,M)
60*      AI=ATAN(TAN(THTA)*SQRT(1+P))
61*      AI=AI/PI

```

TABLE D-1. - Continued.

```

62*      CALL GAT(PGL,PGR,I,J,AI)
63*      SIG=10.**((SIGMA/10.))
64*      IF(J.EQ.1)PGR=0.
65*      SUM=SUM+((PGL+PGR)*SIG*DA/RR)
66*      C      WRITE(6,6000)PGL,PGR,SIGMA,SIG,DA,RR,SUM
67*      400    CONTINUE
68*      C CALCULATE POWER OUTPUT FOR AN INCIDENT ANGLE.
69*      PAC=SUM*CLAMD/CPI/DF
70*      PAC=10.*ALOG10(PAC)
71*      POW(I)=PAC-ALOSS(M)-ROFF(I)+SCAL(M)
72*      C      WRITE(6,6000)PAC,FDOP,DF,CLAMD,CPI,ALOSS(M),SCAL(M)
73*      500    CONTINUE
74*      WRITE(6,3000)IFIL,NS,M
75*      WRITE(6,1000)
76*      WRITE(6,2000)(POW(I),I=1,8)
77*      IFIL=IFIL+1
78*      IF(IFIL.LT.IEOF)GO TO 50
79*      600    CONTINUE
80*      STOP
81*      1000   FORMAT(2H , 'INCIDENT ANGLE AT',3X,'5',11X,'10',10X,'15',10X,'20',
82*      C      10X,'30',10X,'40',10X,'50',10X,'60',/)
83*      2000   FORMAT(16X,8D12.4)
84*      3000   FORMAT(3I10)
85*      4000   FORMAT(22F6.1)
86*      5000   FORMAT(20F6.1)
87*      6000   FORMAT(7D12.4)
88*      END

```

```

1*      SUBROUTINE GAT(GL,GR,I,J,AI)
2*      COMMON/B/IDADD,IX,IY,IXMAX,IYMAX
3*      DIMENSION BUF1(92),BUF2(92)
4*      C DETERMINE THE POSITION OF THE ANTENNA FOOTPRINT.
5*      AI=AI/2.
6*      IA=35-AINT(AI)
7*      AAI=35.-AI
8*      IAP1=IA-1
9*      IP1=IA*IX+IDADD
10*      IP2=(IA-1)*IX+IDADD
11*      C READ ANTENNA RECORDS.
12*      CALL RREAD(IP1,BUF1,IX,IST)
13*      5      IF(IST.EQ.1)GO TO 5
14*      IF(IST.NE.0)WRITE(6,1000)IA
15*      CALL RREAD(IP2,BUF2,IX,IST)
16*      6      IF(IST.EQ.1)GO TO 6
17*      IF(IST.NE.0)WRITE(6,1000)IAP1
18*      C PERFORM LINEAR INTERPOLATION.
19*      IF(J.EQ.1)GO TO 7
20*      PL1=BUF1(47-J)
21*      PL2=BUF2(47-J)

```

TABLE D-1. - Concluded.

```

22*      PR1=BUF1(45+J)
23*      PR2=BUF2(45+J)
24*      PL1L=10.** (PL1/10.)
25*      IF (J.GT.5.AND.PL1.EQ.0.) PL1L=0.
26*      PL2L=10.** (PL2/10.)
27*      IF (J.GT.5.AND.PL2.EQ.0.) PL2L=0.
28*      PR1L=10.** (PR1/10.)
29*      IF (J.GT.5.AND.PR1.EQ.0.) PR1L=0.
30*      PR2L=10.** (PR2/10.)
31*      IF (J.GT.5.AND.PR2.EQ.0.) PR2L=0.
32*      GL=(( (AAI-IA)/(IAP1-IA)) * (PL2L-PL1L)) + PL1L
33*      GR=(( (AAI-IA)/(IAP1-IA)) * (PR2L-PR1L)) + PR1L
34*      GO TO 8
35*      7 P1=BUF1(46)
36*      P2=BUF2(46)
37*      P1=10.** (P1/10.)
38*      P2=10.** (P2/10.)
39*      GL=(( (AAI-IA)/(IAP1-IA)) * (P2-P1)) + P1
40*      GR=GL
41*      8 CONTINUE
42*      C WRITE(6,4000) P1,P2,PL1,PL2,PR1,PR2,GL,GR,AAI,IA,IAP1
43*      RETURN
44*      1000 FORMAT(1H , 'DISC READ PROBLEM FOR ANGLE ',I5)
45*      2000 FORMAT(1H ,F6.2,5X,3I10)
46*      3000 FORMAT(1H ,7F12.4)
47*      4000 FORMAT(1H ,9F8.2,2I10)
48*      END

```

```

1*      SUBROUTINE BCOFF (ANGLE,SIGMA,N,M)
2*      SIGMA=2.10-(0.105*ANGLE)
3*      IF (M.EQ.2.OR.M.EQ.3) GO TO 10
4*      IF (N.EQ.2.OR.N.EQ.3) SIGMA=SIGMA-10.
5*      RETURN
6*      10 CONTINUE
7*      IF (N.EQ.1.OR.N.EQ.4) SIGMA=SIGMA-10.
8*      RETURN
9*      END

```